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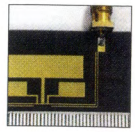

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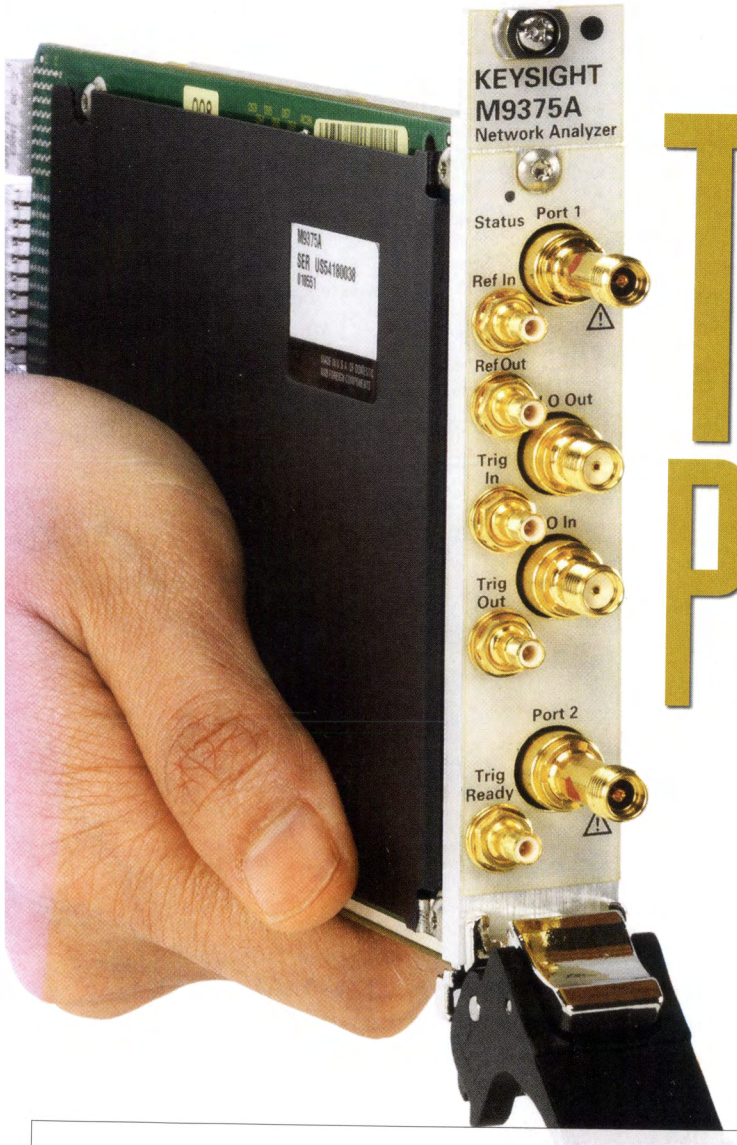
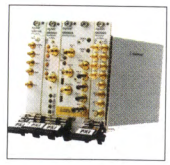
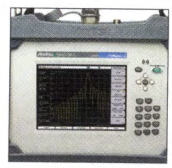
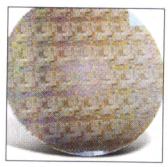


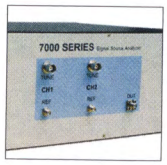
TOP PRODUCTS OF 2014



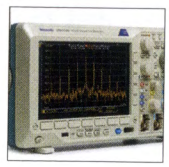


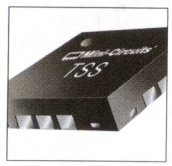


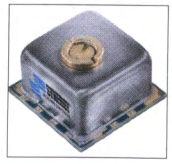












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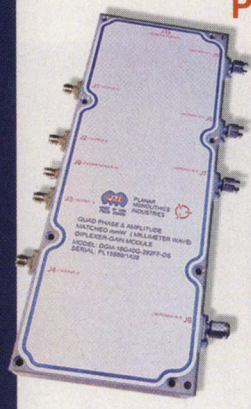

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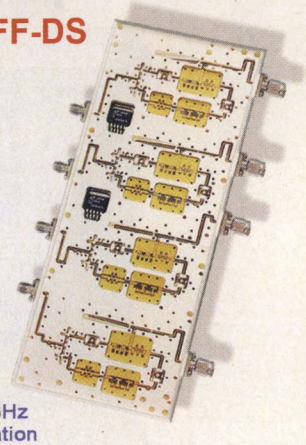
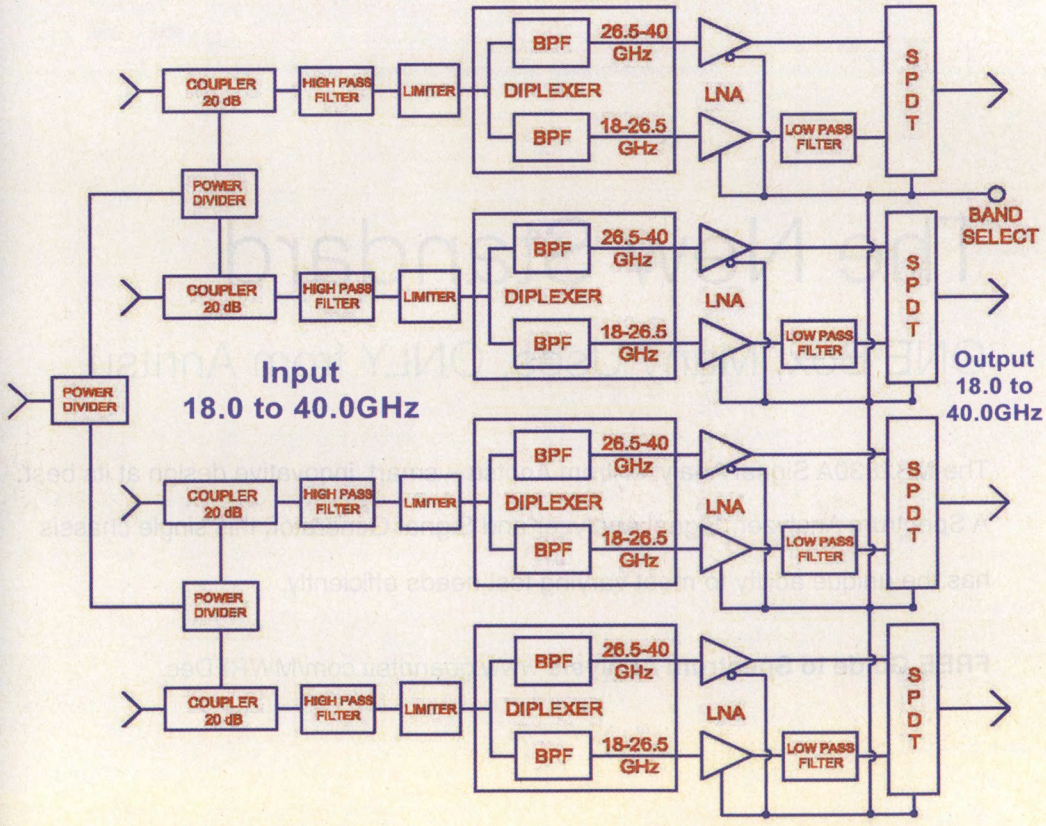


Functional Block Diagram



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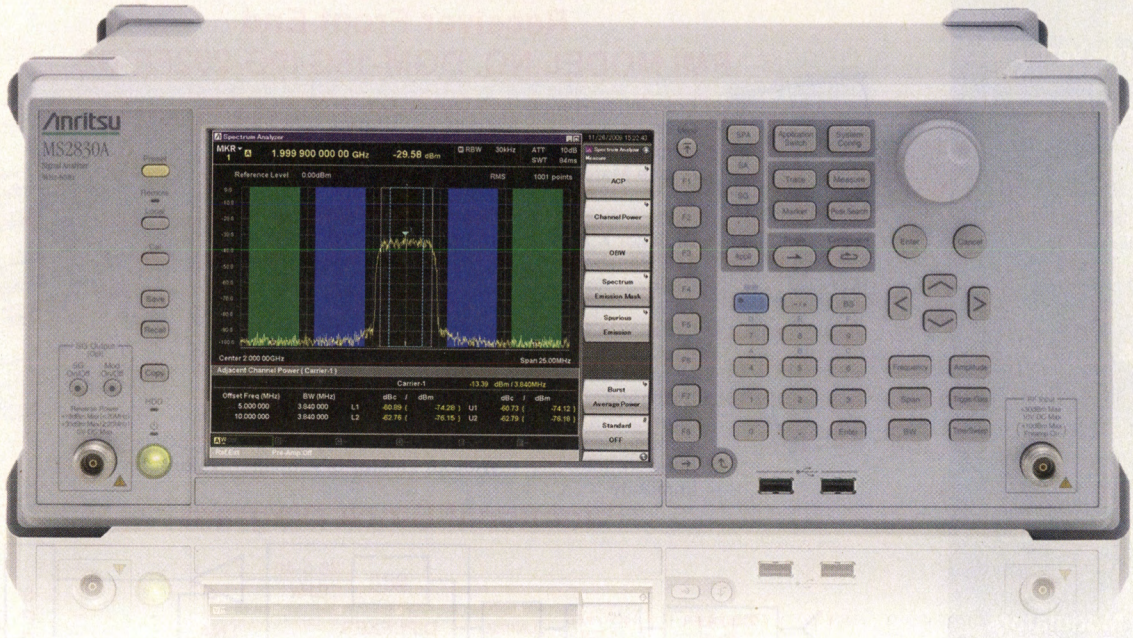

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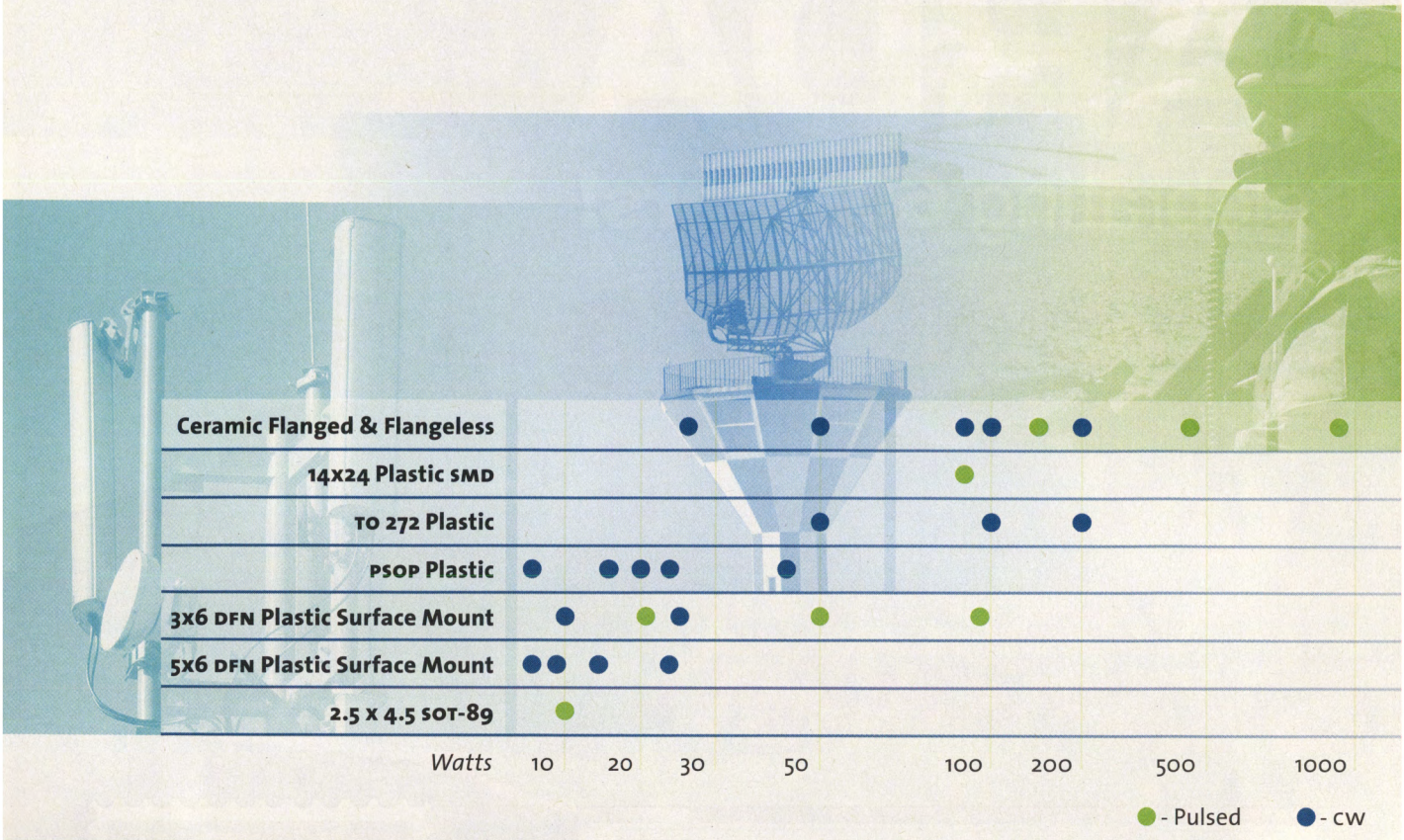

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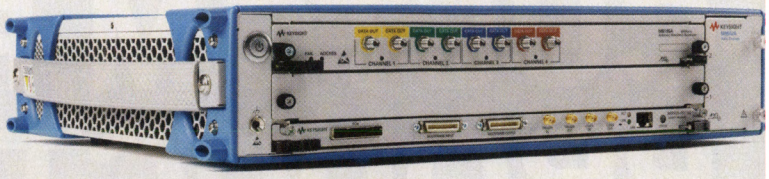
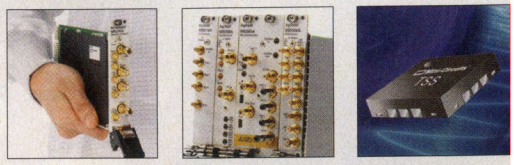


98 **SPECIAL FEATURE:**
AWG DELIVERS 20 GHZ WITH 65 GSAMPLES/S
This arbitrary waveform generator offers 8-b vertical resolution and as many as four synchronized channels to produce complex modulated signals.

98

102 TOP PRODUCTS OF 2014
Product advances during the year benefit customers of active and passive components, as well as test equipment.



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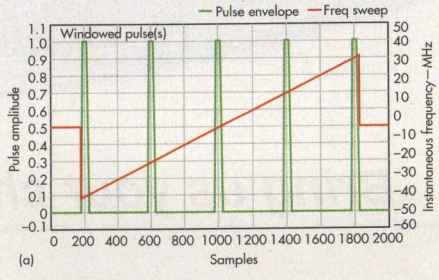
55 THE EVOLUTION OF HARMONIC LOAD PULL
Harmonic load-pull tuners are important systems for characterizing power transistors and amplifiers.

64 BANDPASS FILTER COMBINES HIGH SELECTIVITY, REJECTION
By using a stepped-impedance stub-loaded multimode resonator, this wideband filter achieves both high selectivity and good out-of-band rejection.

72 LOW-NOISE VCO RUNS ON LOW VOLTAGE
This quadrature PMOS Colpitts VCO requires very little power for stable, low-phase-noise operation at 2.4 GHz.

80 SIMULATE AND TEST CHIRPED RADAR
Frequency-stepped chirped radar systems are being used in automotive, law-enforcement, and public-transportation applications.

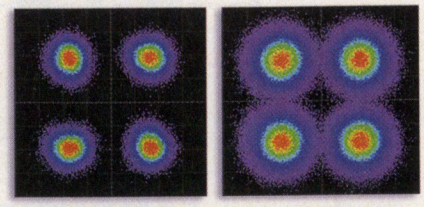


80

84 LNA INTEGRATES FAST SHUTDOWN FUNCTION
This low-noise amplifier includes an on-board switching function with industry-leading switching speed.

90 WILKINSON DIVIDER COVERS TWO BANDS
This novel, low-loss power divider covers two frequency bands with a straightforward design.





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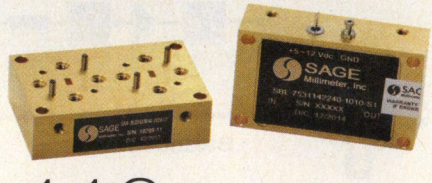

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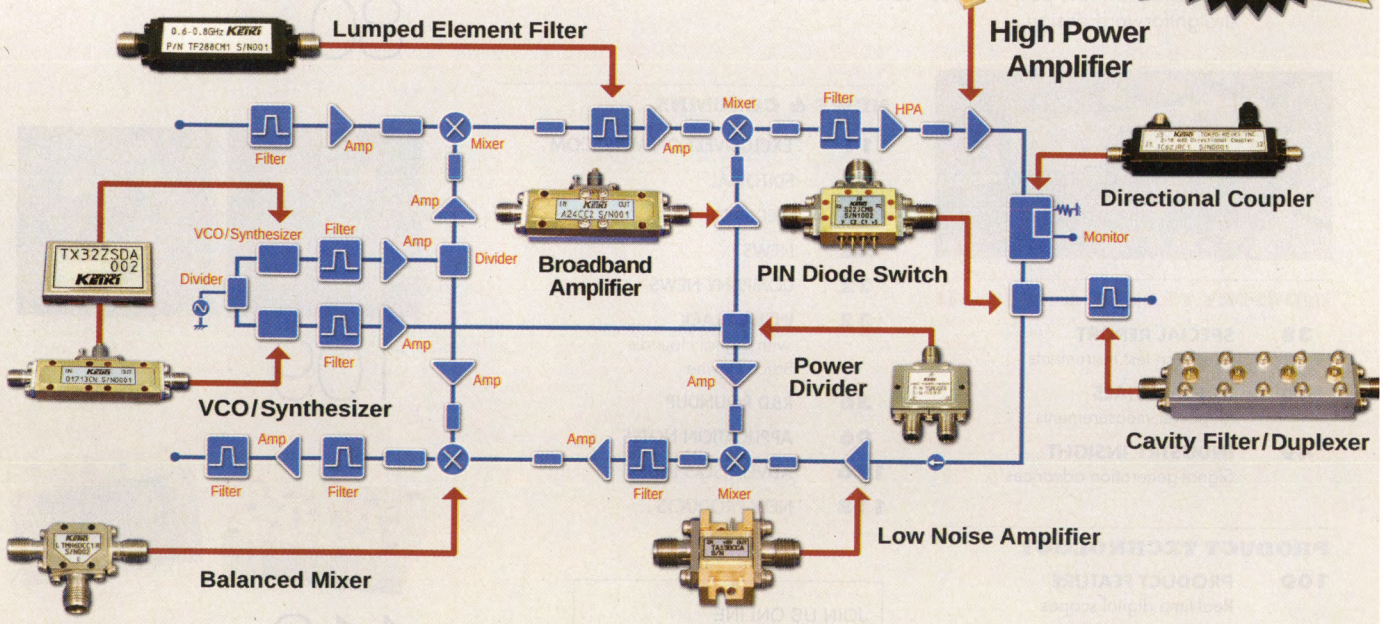

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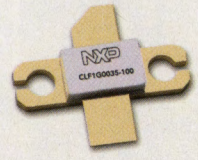



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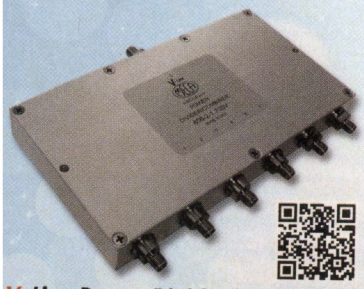

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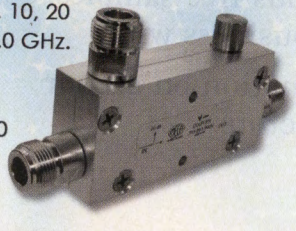

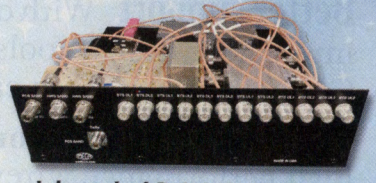

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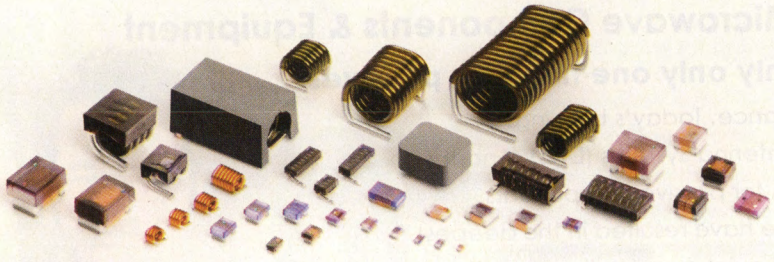

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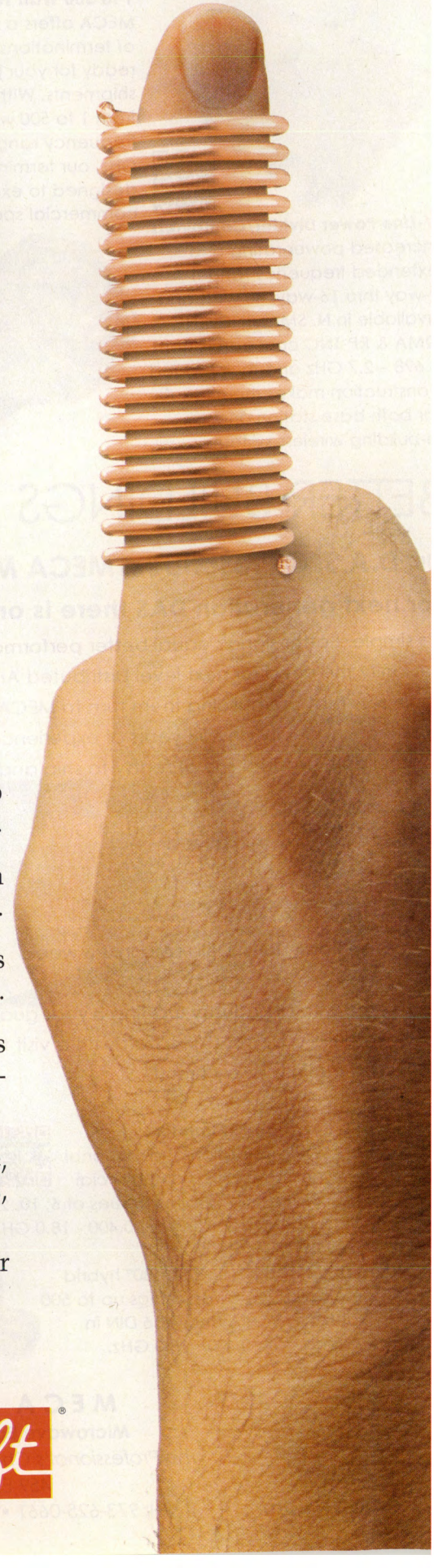

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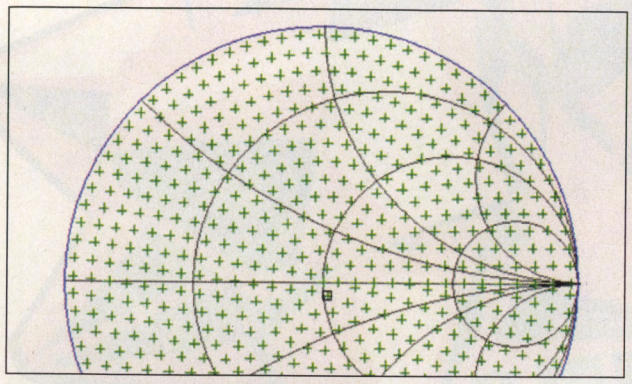

IMPEDANCE TUNING 101

http://mwrf.com/test-measurement/impedance-tuning-101

Non-50ohm measurements have become a hot topic over the past few years, but so much information might seem overwhelming for a beginner. Here's a basic introduction to non-50ohm measurement concepts and the need for such measurements from Gary Simpson, chief tech officer at Maury Microwave—who also is the subject of an online Q&A this month.

VIDEO: PRECISION 150 MM PROBE SYSTEM FOR MMW

http://mwrf.com/cascade-microtech-cost-effective-precision-150-mm-probe-system-mmw

Microwaves & RF's Nancy Friedrich chats with Thomas Tharigen, Cascade Microtech product marketing manager, about the EPS150MMW, a dedicated manual probing solution for achieving accurate measurement results in the shortest time. The backlash-free, high-precision X-Y-Z movement of mmW and RF positioners, integrated planarization, and a contact separation drive with 1 µm repeatability, enables precise probe placement and contact repeatability comparable to semi-automatic systems.



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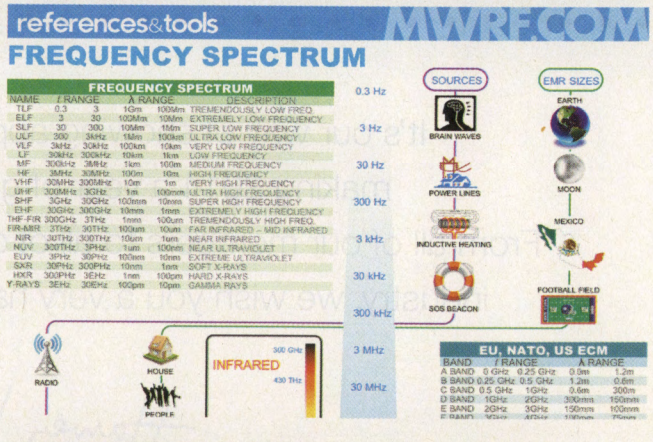

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Editorial

NANCY K. FRIEDRICH
Content Director
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Get with the Times, Barbie

It's the holidays—a time when all of us with little kids in our lives can choose gifts that promote science, technology, engineering, and math (STEM). Today, there is a growing number of wonderful toys, games, books, movies, and activities that promote STEM. In picture books alone, recent titles have included *Rosie Revere, Engineer*, and *The Most Magnificent Thing*. Just make sure you avoid the Barbie book from 2010 titled *Barbie: I Can Be A Computer Engineer*.

In the book, Barbie is supposedly working to create a computer program. But then she clarifies that she is "only creating the design ideas" because she'll need "Steven's and Brian's help to turn it into a real game." When she encounters problems in the process, she enlists the help of the two male programmers, who of course save the day.

Considering the book's negative message, Mattel has pulled it from the shelves. But the company didn't act in time to avoid bad press and, in this age of social media, quite a fun rebirth. Kathleen Tuite, a computer programmer and the daughter of Don Tuite (an editor on our sister publication, *Electronic Design*), created a website that allowed people to re-caption the book. People then took to Twitter to promote the new versions or comment on them using the #FeministHackerBarbie hashtag, which makes them easy to find and enjoy.

While the tech community quickly engaged in correcting the book's negative image, many troubling stereotypes and attitudes still exist around girls—and kids in general—being "uncool" if they're smart. That type of thinking should change as effort continues to be invested in getting children—and young girls in particular—into STEM fields. After all, attracting the next generation of engineers is already proving a challenge.

So if you're buying a present for a kid in your life, think about something that encourages STEM interests. Talk to kids about what you do. Share your excitement and enthusiasm. Today's kids are our future engineers. Given the right knowledge, experience, and passion, it's impossible to fathom where they will take today's technology. From all of us at *Microwaves & RF*, may you have a wonderful holiday and a healthy and happy New Year! MW

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LS00120 P10A	10 - 2000	0.50		
LS00130 P10A	10 - 3000	0.60	+14	+18
LS00140 P10A	10 - 4000	0.70		
LS00160 P10A	10 - 6000	1.30		
LS00210 P10A	20 - 1000	0.40		
LS00220 P10A	20 - 2000	0.50		
LS00230 P10A	20 - 3000	0.60	+14	+18
LS00240 P10A	20 - 4000	0.70		
LS00260 P10A	20 - 6000	1.30		
LS00510 P10A	50 - 1000	0.40		
LS00520 P10A	50 - 2000	0.50		
LS00530 P10A	50 - 3000	0.60	+14	+18
LS00540 P10A	50 - 4000	0.70		
LS00560 P10A	50 - 6000	1.20		

Note 1. Insertion Loss and VSWR (typically 1.2:1 to 1.4:1) tested at -10 dBm.
Note 2. Typical limiting threshold: +6 dBm
Note 3. Power rating derated to 20% @ +125 Deg. C.
Note 4. Leakage +16 dBm maximum for 1 W CW input, +21 dBm for 10W CW input for frequencies below 100 MHz.

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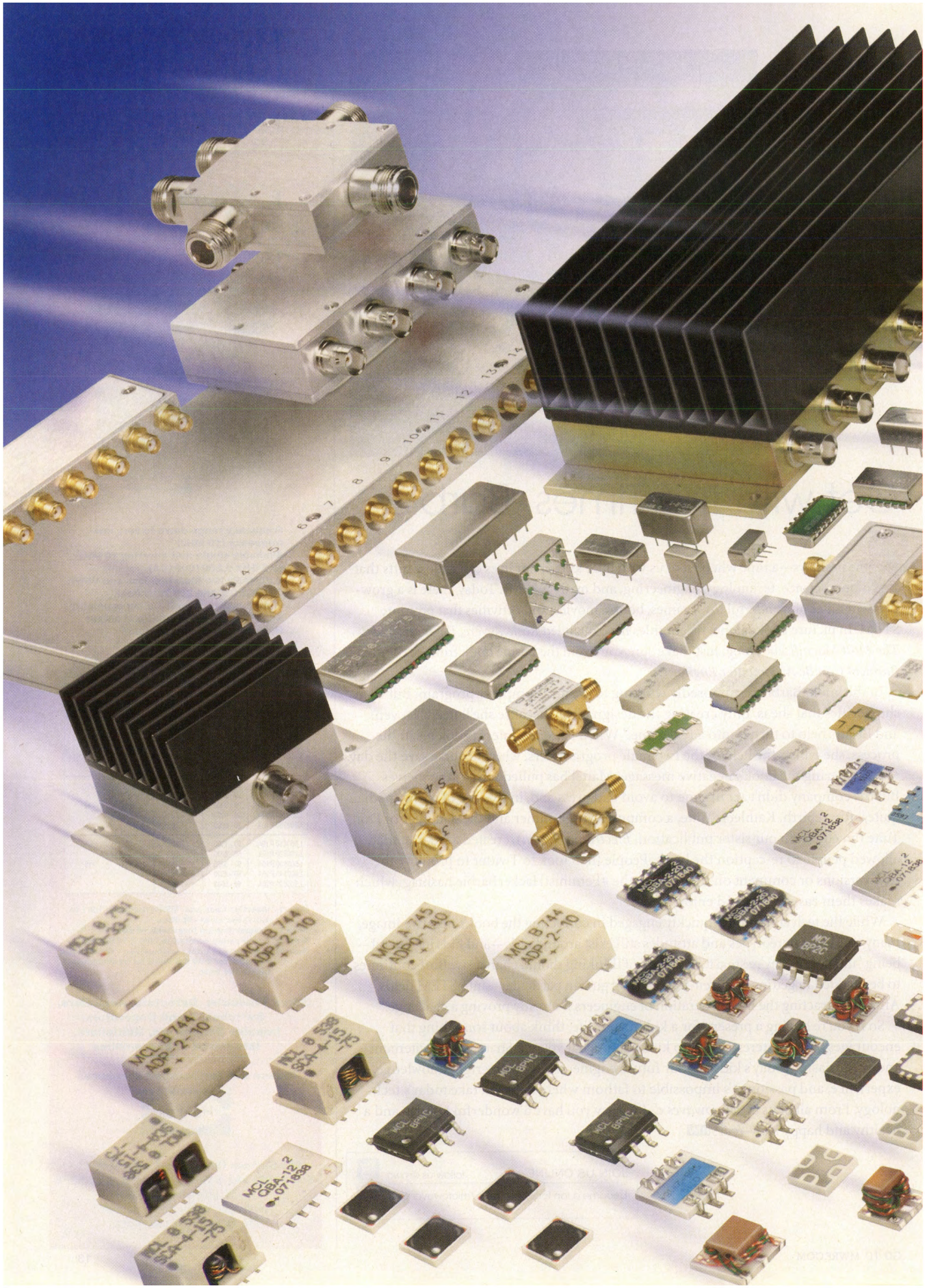

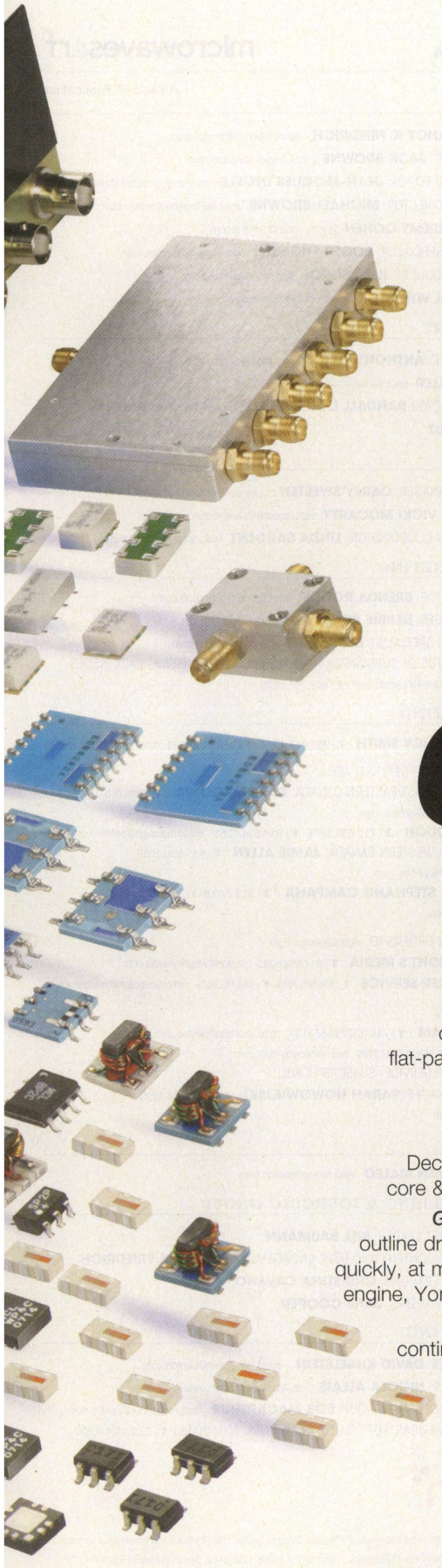

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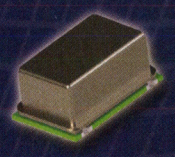

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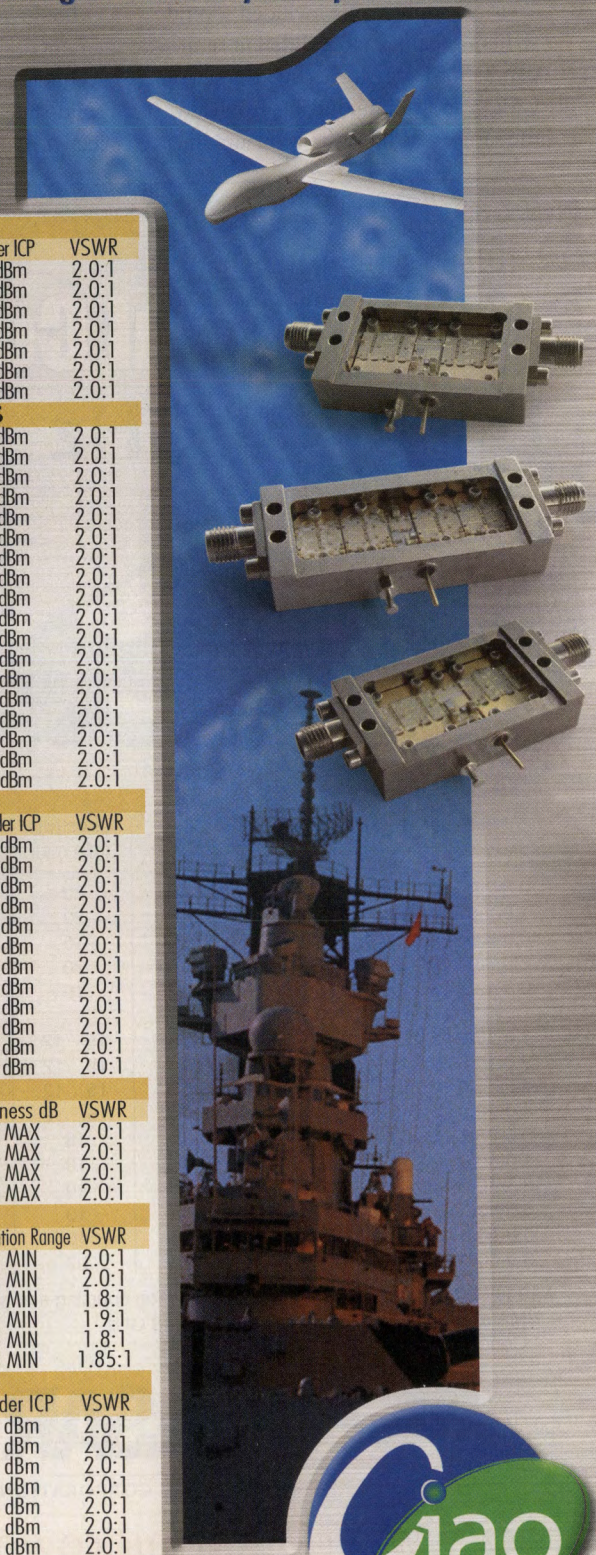

OCTAVE BAND LOW NOISE AMPLIFIERS

Model No.	Freq (GHz)	Gain (dB) MIN	Noise Figure (dB)	Power-out @ P1-dB	3rd Order ICP	VSWR
CA01-2110	0.5-1.0	28	1.0 MAX, 0.7 TYP	+10 MIN	+20 dBm	2.0:1
CA12-2110	1.0-2.0	30	1.0 MAX, 0.7 TYP	+10 MIN	+20 dBm	2.0:1
CA24-2111	2.0-4.0	29	1.1 MAX, 0.95 TYP	+10 MIN	+20 dBm	2.0:1
CA48-2111	4.0-8.0	29	1.3 MAX, 1.0 TYP	+10 MIN	+20 dBm	2.0:1
CA812-3111	8.0-12.0	27	1.6 MAX, 1.4 TYP	+10 MIN	+20 dBm	2.0:1
CA1218-4111	12.0-18.0	25	1.9 MAX, 1.7 TYP	+10 MIN	+20 dBm	2.0:1
CA1826-2110	18.0-26.5	32	3.0 MAX, 2.5 TYP	+10 MIN	+20 dBm	2.0:1

NARROW BAND LOW NOISE AND MEDIUM POWER AMPLIFIERS

Model No.	Freq (GHz)	Gain (dB) MIN	Noise Figure (dB)	Power-out @ P1-dB	3rd Order ICP	VSWR
CA01-2111	0.4 - 0.5	28	0.6 MAX, 0.4 TYP	+10 MIN	+20 dBm	2.0:1
CA01-2113	0.8 - 1.0	28	0.6 MAX, 0.4 TYP	+10 MIN	+20 dBm	2.0:1
CA12-3117	1.2 - 1.6	25	0.6 MAX, 0.4 TYP	+10 MIN	+20 dBm	2.0:1
CA23-3111	2.2 - 2.4	30	0.6 MAX, 0.45 TYP	+10 MIN	+20 dBm	2.0:1
CA23-3116	2.7 - 2.9	29	0.7 MAX, 0.5 TYP	+10 MIN	+20 dBm	2.0:1
CA34-2110	3.7 - 4.2	28	1.0 MAX, 0.5 TYP	+10 MIN	+20 dBm	2.0:1
CA56-3110	5.4 - 5.9	40	1.0 MAX, 0.5 TYP	+10 MIN	+20 dBm	2.0:1
CA78-4110	7.25 - 7.75	32	1.2 MAX, 1.0 TYP	+10 MIN	+20 dBm	2.0:1
CA910-3110	9.0 - 10.6	25	1.4 MAX, 1.2 TYP	+10 MIN	+20 dBm	2.0:1
CA1315-3110	13.75 - 15.4	25	1.6 MAX, 1.4 TYP	+10 MIN	+20 dBm	2.0:1
CA12-3114	1.35 - 1.85	30	4.0 MAX, 3.0 TYP	+33 MIN	+41 dBm	2.0:1
CA34-6116	3.1 - 3.5	40	4.5 MAX, 3.5 TYP	+35 MIN	+43 dBm	2.0:1
CA56-5114	5.9 - 6.4	30	5.0 MAX, 4.0 TYP	+30 MIN	+40 dBm	2.0:1
CA812-6115	8.0 - 12.0	30	4.5 MAX, 3.5 TYP	+30 MIN	+40 dBm	2.0:1
CA812-6116	8.0 - 12.0	30	5.0 MAX, 4.0 TYP	+33 MIN	+41 dBm	2.0:1
CA1213-7110	12.2 - 13.25	28	6.0 MAX, 5.5 TYP	+33 MIN	+42 dBm	2.0:1
CA1415-7110	14.0 - 15.0	30	5.0 MAX, 4.0 TYP	+30 MIN	+40 dBm	2.0:1
CA1722-4110	17.0 - 22.0	25	3.5 MAX, 2.8 TYP	+21 MIN	+31 dBm	2.0:1

ULTRA-BROADBAND & MULTI-OCTAVE BAND AMPLIFIERS

Model No.	Freq (GHz)	Gain (dB) MIN	Noise Figure (dB)	Power-out @ P1-dB	3rd Order ICP	VSWR
CA0102-3111	0.1-2.0	28	1.6 Max, 1.2 TYP	+10 MIN	+20 dBm	2.0:1
CA0106-3111	0.1-6.0	28	1.9 Max, 1.5 TYP	+10 MIN	+20 dBm	2.0:1
CA0108-3110	0.1-8.0	26	2.2 Max, 1.8 TYP	+10 MIN	+20 dBm	2.0:1
CA0108-4112	0.1-8.0	32	3.0 MAX, 1.8 TYP	+22 MIN	+32 dBm	2.0:1
CA02-3112	0.5-2.0	36	4.5 MAX, 2.5 TYP	+30 MIN	+40 dBm	2.0:1
CA26-3110	2.0-6.0	26	2.0 MAX, 1.5 TYP	+10 MIN	+20 dBm	2.0:1
CA26-4114	2.0-6.0	22	5.0 MAX, 3.5 TYP	+30 MIN	+40 dBm	2.0:1
CA618-4112	6.0-18.0	25	5.0 MAX, 3.5 TYP	+23 MIN	+33 dBm	2.0:1
CA618-6114	6.0-18.0	35	5.0 MAX, 3.5 TYP	+30 MIN	+40 dBm	2.0:1
CA218-4116	2.0-18.0	30	3.5 MAX, 2.8 TYP	+10 MIN	+20 dBm	2.0:1
CA218-4110	2.0-18.0	30	5.0 MAX, 3.5 TYP	+20 MIN	+30 dBm	2.0:1
CA218-4112	2.0-18.0	29	5.0 MAX, 3.5 TYP	+24 MIN	+34 dBm	2.0:1

LIMITING AMPLIFIERS

Model No.	Freq (GHz)	Input Dynamic Range	Output Power Range Psat	Power Flatness dB	VSWR
CLA24-4001	2.0 - 4.0	-28 to +10 dBm	+7 to +11 dBm	+/- 1.5 MAX	2.0:1
CLA26-8001	2.0 - 6.0	-50 to +20 dBm	+14 to +18 dBm	+/- 1.5 MAX	2.0:1
CLA712-5001	7.0 - 12.4	-21 to +10 dBm	+14 to +19 dBm	+/- 1.5 MAX	2.0:1
CLA618-1201	6.0 - 18.0	-50 to +20 dBm	+14 to +19 dBm	+/- 1.5 MAX	2.0:1

AMPLIFIERS WITH INTEGRATED GAIN ATTENUATION

Model No.	Freq (GHz)	Gain (dB) MIN	Noise Figure (dB)	Power-out @ P1-dB	Gain Attenuation Range	VSWR
CA001-2511A	0.025-0.150	21	5.0 MAX, 3.5 TYP	+12 MIN	30 dB MIN	2.0:1
CA05-3110A	0.5-5.5	23	2.5 MAX, 1.5 TYP	+18 MIN	20 dB MIN	2.0:1
CA56-3110A	5.85-6.425	28	2.5 MAX, 1.5 TYP	+16 MIN	22 dB MIN	1.8:1
CA612-4110A	6.0-12.0	24	2.5 MAX, 1.5 TYP	+12 MIN	15 dB MIN	1.9:1
CA1315-4110A	13.75-15.4	25	2.2 MAX, 1.6 TYP	+16 MIN	20 dB MIN	1.8:1
CA1518-4110A	15.0-18.0	30	3.0 MAX, 2.0 TYP	+18 MIN	20 dB MIN	1.85:1

LOW FREQUENCY AMPLIFIERS

Model No.	Freq (GHz)	Gain (dB) MIN	Noise Figure dB	Power-out @ P1-dB	3rd Order ICP	VSWR
CA001-2110	0.01-0.10	18	4.0 MAX, 2.2 TYP	+10 MIN	+20 dBm	2.0:1
CA001-2211	0.04-0.15	24	3.5 MAX, 2.2 TYP	+13 MIN	+23 dBm	2.0:1
CA001-2215	0.04-0.15	23	4.0 MAX, 2.2 TYP	+23 MIN	+33 dBm	2.0:1
CA001-3113	0.01-1.0	28	4.0 MAX, 2.8 TYP	+17 MIN	+27 dBm	2.0:1
CA002-3114	0.01-2.0	27	4.0 MAX, 2.8 TYP	+20 MIN	+30 dBm	2.0:1
CA003-3116	0.01-3.0	18	4.0 MAX, 2.8 TYP	+25 MIN	+35 dBm	2.0:1
CA004-3112	0.01-4.0	32	4.0 MAX, 2.8 TYP	+15 MIN	+25 dBm	2.0:1

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FINDING SOLUTIONS THROUGH CURIOSITY

Your November issue included several writers debating about different semiconductor technologies, such as silicon CMOS and gallium nitride. Ironically, the same issue had a story on amplifiers using electron tubes, one of the older active device technologies still in use.

Lower-frequency (audio) electronic circuit designers have often favored older technologies, such as electron tubes, over solid-state-device amplification. The subjective opinions of listeners at those audio frequencies have supported superior sonic performance from electron tubes compared to solid-state devices. Still, the RF/microwave industry supplies active devices for many applications that must operate with the most spectrally accurate and distortion-free devices available. In most cases, solid-state devices operate with considerably less distortion than electron-tube devices.

Granted, some applications may require high-power signal levels that can be difficult for solid-state devices to achieve and maintain. Higher power levels at microwave frequencies can be reached by teaming multiple solid-state devices with power-combining circuits, but individual solid-state devices are still no match for the power levels possible from individual electron-tube devices.

But high-frequency researchers are a curious bunch. This same curiosity has encouraged the use of novel materials over the years like indium phosphide (InP) and gallium nitride (GaN). There no doubt will be other materials to be explored for use as high-frequency, high-power semiconductors, as researchers search for better-performing, lower-costing high-frequency solutions. Consider that gallium arsenide (GaAs) was once considered the ultimate high-frequency semiconductor material by researchers and, during that same timeframe, GaN was essentially unknown.

For all of the DARPA funds that have been invested in finding higher-power GaAs device solutions over the last few decades, GaAs power devices are now viewed more as historical curiosities than anything else when compared to the capabilities of GaN solid-state power devices. And for all of the excitement connected to GaN technology for its high-frequency, high-power potential and capabilities, it should not be so surprising to look back 10 years from now and wonder how so much hope could have been placed in GaN when it has been replaced by the next great thing.

Frank Kaminsky

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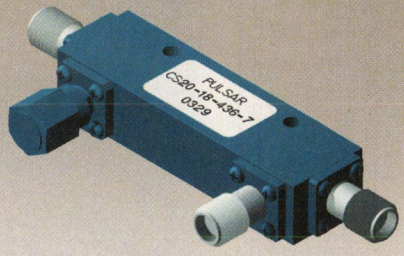

Frequency Range	I.L. (dB) min.	Coupling Flatness max.	Directivity (dB) min.	VSWR max.	Model Number
0.5-2.0 GHz	0.35	± 0.75 dB	23	1.20:1	CS*-02
1.0-4.0 GHz	0.35	± 0.50 dB	23	1.20:1	CS*-04
0.5-6.0 GHz	1.00	± 0.80 dB	15	1.50:1	CS10-24
2.0-8.0 GHz	0.35	± 0.40 dB	20	1.25:1	CS*-09
0.5-12.0 GHz	1.00	± 0.80 dB	15	1.50:1	CS*-19
1.0-18.0 GHz	0.90	± 0.50 dB	15 12	1.50:1	CS*-18
2.0-18.0 GHz	0.80	± 0.50 dB	15 12	1.50:1	CS*-15
4.0-18.0 GHz	0.60	± 0.50 dB	15 12	1.40:1	CS*-16
8.0-20.0 GHz	1.00	± 0.80 dB	15 12	1.50:1	CS*-21
6.0-26.5 GHz	0.70	± 0.80 dB	13	1.55:1	CS20-50
1.0-40.0 GHz	1.60	± 1.50 dB	10	1.80:1	CS20-53
2.0-40.0 GHz	1.60	± 1.00 dB	10	1.80:1	CS20-52
6.0-40.0 GHz	1.20	± 1.00 dB	10	1.70:1	CS10-51
6.0-50.0 GHz	1.60	± 1.00 dB	10	2.00:1	CS20-54
6.0-60.0 GHz	1.80	± 1.00 dB	07	2.00:1	CS20-55

10 to 500 watts power handling depending on coupling and model number..
SMA and Type N connectors available to 18 GHz.
* Coupling Value: 3, 6, 8, 10, 13, 16, 20 dB.

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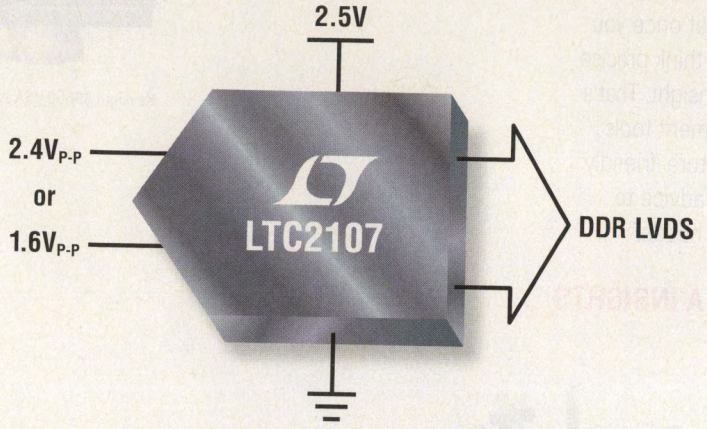

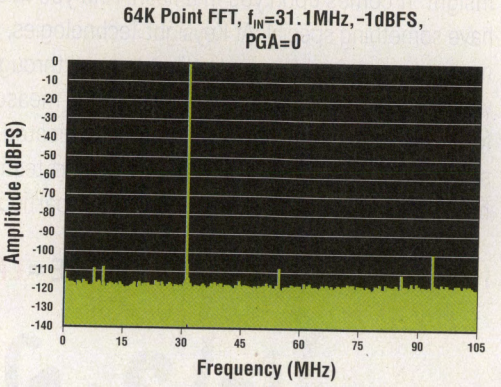

LTC®2107 Achieves 80dB SNR from Single 2.5V Supply

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- 100dBFS SFDR
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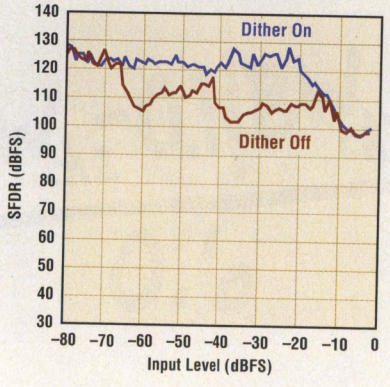
- CMOS or DDR/LVDS Outputs
- Single 2.5V Supply
- Clock Duty Cycle Stabilizer
- 7mm x 7mm QFN Package

SFDR vs Input Level, fIN=30.6MHz, PGA=0



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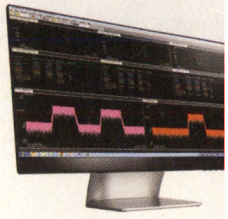

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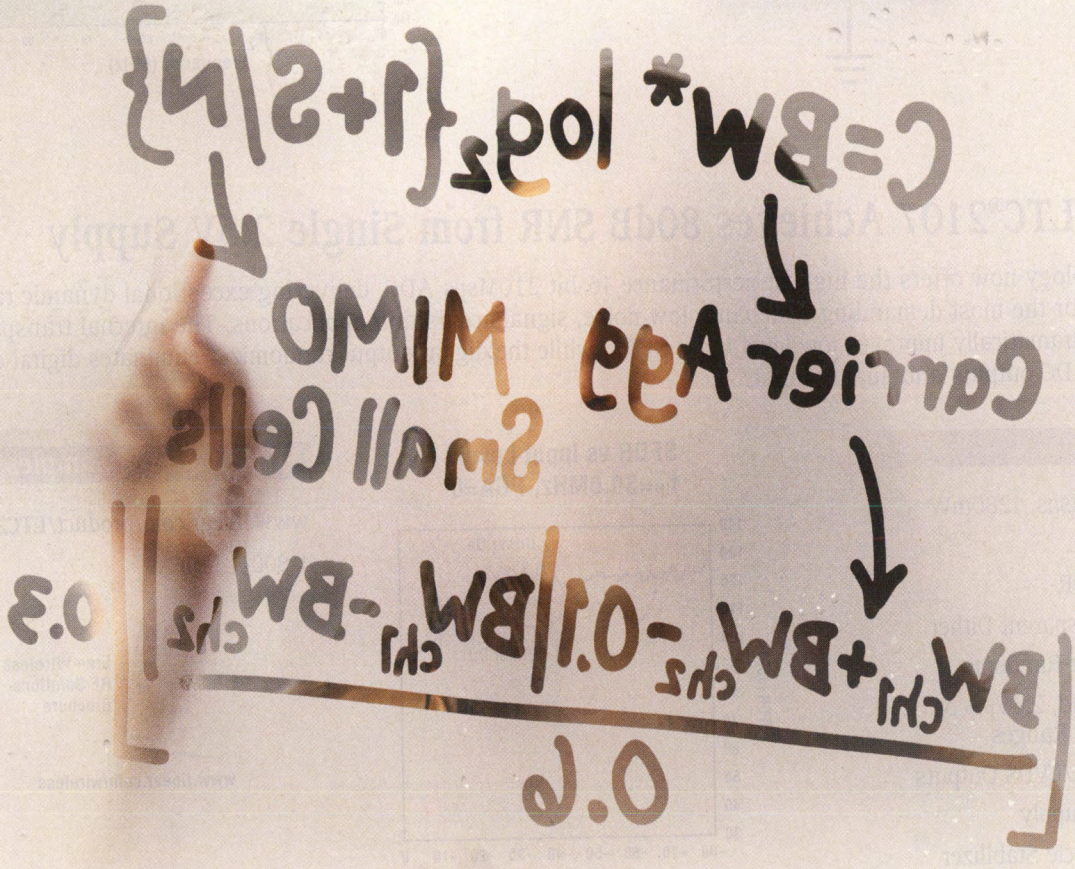

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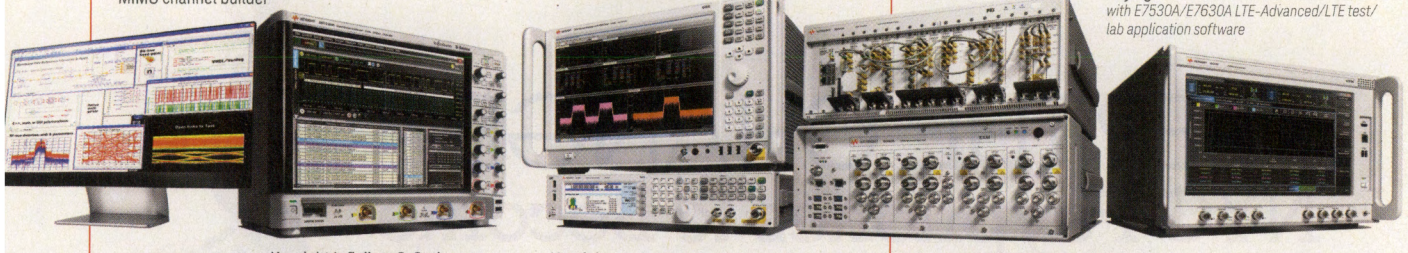

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News



KEEP THE CHANGE—
NYC Payphones to Become Free Wi-Fi Stations

As part of an effort to bring municipal Wi-Fi to the millions of people traveling through and living in New York City, the LinkNYC network aims to transform outdated payphones into state-of-the-art wireless connection points. The network, which will span across all five boroughs, will provide free 24/7 Internet access at speeds up to 1 Gbit/s, free phone calls to anywhere in the U.S., easy access to 911 and 211, and a free charging station for mobile devices. It's expected the stations will generate more than $500 million in revenue for the city over the next 12 years.

The encrypted network connection will be implemented by the CityBridge team behind LinkNYC, a consortium of members including Comark, Control Group, Qualcomm, and Titan. The stations will feature an Android touchscreen interface to access city services and directions alongside digital displays for advertising and public service announcements. Up to 10,000 Link stations will be installed (beginning in 2015 with operation expected by the end of the year), delivering Wi-Fi speeds that are 100 times faster than conventional municipal Wi-Fi.

The amount of installed Link stations will also create a seamless roaming experience, thanks to the Wi-Fi signals within the city's subway stations provided by Transit Wireless (it will also integrate the high-speed fiber infrastructure to power the new stations). The signals, however, will only work within 150 ft. of each 9.5 ft. tall hub. The company's fiber-optic cables currently run through ducts under the city streets to the subway stations to connect the multi-band RF nodes within each station to distribute the signals.



Next year, payphones in New York City will begin turning into Wi-Fi hubs reaching speeds up to 1 Gbit/s.
(Image courtesy of CityBridge)

Transit Wireless' Wi-Fi network runs parallel to the cellular distributed antenna system (DAS), providing a fiber-to-edge network that aims to maximize throughput and performance. Access points deliver 2.4- and 5-GHz public unlicensed band, and 4.9-GHz public safety band, connectivity throughout locations in each station. Similar technology will be used to provide the network for the Link stations.

According to CNET, LinkNYC comes despite a mixed history of municipal Wi-Fi projects, with cities throughout the U.S. trying—and failing—to create and sustain such services. Many of these projects were scrapped or scaled back over the years after cities and developers struggled with financial, competitive ,and technical challenges. Several current municipal Wi-Fi networks are available only in downtown areas and for a few hours a day for users.

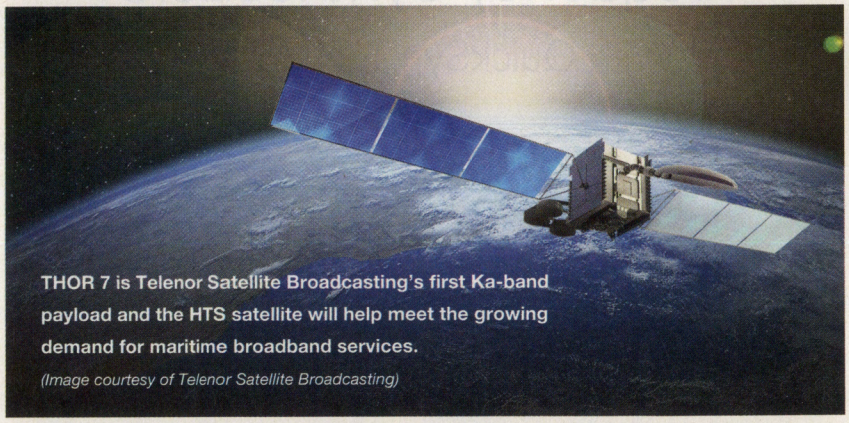
The CityBridge team has already piloted a few Wi-Fi payphone hotspots over the past two years, though, and city residents won't have to pay any tax dollars toward construction of the new network. The contracted expenses, in combination with the prime advertising space and subsequent revenue, are expected to push the project onto a successful and maintainable level. ■

HIGH-THROUGHPUT SATELLITES Prepare for Steady Growth

THE INFLUX OF high-throughput-satellite (HTS) capacity and the adoption of satellites using this technology are set to accelerate over the next decade, according to a report from Euroconsult. "High Throughput Satellites: On Course for New Horizons" predicts that HTS systems are expected to grow from 107 GBytes/s in 2014 to just over 1300 GBytes/s in 2023—a CAGR of more than 30% over the said period. HTS capacity supply is projected to nearly triple in the next three years alone, rising from 600 GBytes/s in 2014 to 1720 GBytes/s in 2017.

This growth stems, in part, from the extensive investments made by satellite operators in HTS-capacity supply ranging from Ka-band to more recent Ku-band adoption. Examples of these operators include Telesat, SES, and Eutelsat. In addition, various factors are expected to aid in the expansion of HTS services. While there are currently 2.5 million consumer broadband subscribers, that number is expected to grow to 8.8 million by 2023. The number of commercial satellites is expected to grow from 6 to 22,000 over the same time period. In the next decade, the number of connected mobile 3G satellites will increase from 20 to 6000 while the number of active HTS satellite operators will increase from 14 to 25.

Euroconsult expects North America's consumer broadband services to remain the largest user of HTS capacity through 2023, comprising 35% of total usage. Together with cellular-backhaul trunking, civil-government and enterprise networks are expected to grow in usage (34% by 2023). This growth will be derived from demand in Latin America, the Middle East and Africa, and the Asia-Pacific region. HTS demand is also expected to accelerate in commercial mobility markets—specifically, aviation and maritime. Whereas military satellite-communications (satcom) usage is expected to remain modest through 2023, HTS demand for video services is expected to gradually grow over



THOR 7 is Telenor Satellite Broadcasting's first Ka-band payload and the HTS satellite will help meet the growing demand for maritime broadband services.

(Image courtesy of Telenor Satellite Broadcasting)

RADIO-WAVE CIRCULATOR Frees Telecommunications from Magnetic Hold

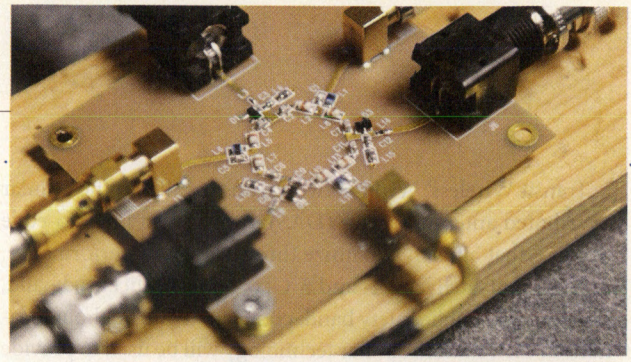
ALTHOUGH MAGNETIC-BASED CIRCULATORS have reigned supreme for providing two-way communications on the same frequency channel, problems arise due to their large size, weight, and cost. To address those issues, researchers at the Cockrell School of Engineering at the University of Texas at Austin, developed a new radio-wave circulator that eliminates the need for magnets and magnetic materials.

Its lack of magnetic materials means a smaller footprint and less expensive materials, suiting the circulator for cellphones and other microelectronic systems. In fact, according to the research team, the circulator could double the useful bandwidth in wireless communications.

Full-duplex functionality—the ability to transmit and receive signals on the same frequency band at the same time—holds the key to its success. The new circulator mimics the way magnetic materials break the symmetry in wave transmission to selectively

route radio waves, but instead of using magnetic bias, a traveling wave spins around the device. It can also be tuned in real time over a broad range of frequencies unlike traditional circulators.

Measuring 2 cm, the prototype device is more than 75% smaller than the wavelength of operation. Purportedly, it could be scaled down to as small as a few microns. The circulator's design is based on commonly used integrated-circuit materials, including gold, copper, and silicon, which makes it easier to integrate into the circuit boards of current communication devices. The circulator's scalability, however, could push its application into phased-array and radar systems for aircraft, ships, and satellites. ■



The circulator's design is based on commonly used integrated-circuit materials, including gold, copper, and silicon, which makes it easier to integrate into the circuit boards of current communication devices. *(Image courtesy of the Cockrell School of Engineering)*

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IC BRIDGES THE "Terahertz Gap" to Break World Record

Researchers persistently try to detect, process, and radiate the high-frequency terahertz section of the electromagnetic spectrum without having to utilize frequency translation or multiplication. Often referred to as the "terahertz gap," a team at Northrop Grumman recently developed what Guinness World Records deemed the world's fasted integrated-circuit amplifier. The amplifier uses 10 transistor stages to reach an operating speed of 1 THz, or one trillion cycles per second.

The terahertz monolithic integrated circuit (TMIC) bridges the gap using a super-scaled, 25-nm gate length, indium-phosphide, high electron mobility transistor. The transistor measures at a gain of 10 dB at 1 THz and 9 dB at 1.03 THz. The Defense Advanced Research Projects Agency (DARPA) funded the research/development of the TMIC. DARPA's goal was to demonstrate tran-

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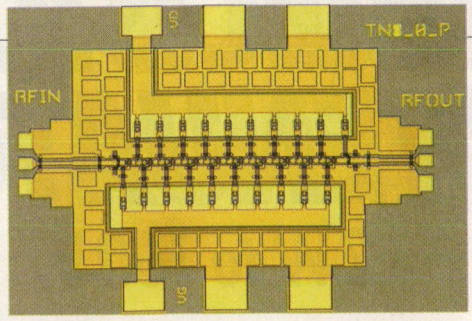
sistor-based electronics operating at 670 GHz, 850 GHz, and 1 THz.

Applications for the technology include high-resolution security imaging systems, improved collision-avoidance radar, communications networks with greater capacity, atmospheric sensing, radio astronomy, and medical imaging. It could also help improve system range and reduce size, weight, and power consumption of existing systems. The amplifier surpasses the company's previous record of 850 billion cycles per second.



The terahertz monolithic integrated circuit (TMIC) bridges the gap using a super-scaled, 25-nm gate length, indium-phosphide, high-electron-mobility transistor. *(Image courtesy of DARPA)*

3D-PRINTED ALUMINUM ANTENNA Support May Be Space-Bound

SIMILAR TO THE consumer world, more industrial manufacturers are catching onto 3D printing. One recent example arose from a collaborative effort that leveraged additive manufacturing to enhance an Earth observation satellite antenna.

A collaboration between Altair, EOS, and RUAG Space utilized topology optimization and 3D printing to create an aluminum support that could achieve optimized weight and stiffness performance characteristics not possible through conventional manufacturing techniques. The component is currently undergoing intensive qualification tests.

Design tools used in the project included Altair's HyperWorks solver Optistruct and solidThinking Evolve to help guarantee a load-sufficient material distribution and meet the necessary design parameters in four weeks. EOS handled the manufacturing of the aluminum components, with the antenna about 40 cm long—reportedly one of the longest metal components ever produced using the powder-bed manufacturing method.

To construct the metal 3D print, built-up layers of powder were joined together to form the desired shape via automated laser sintering. The process results in a much lighter component, which provides satellite engineers with greater design freedom and lowers costs.

The team hopes that the antenna support will be equipped on future Sentinel-1



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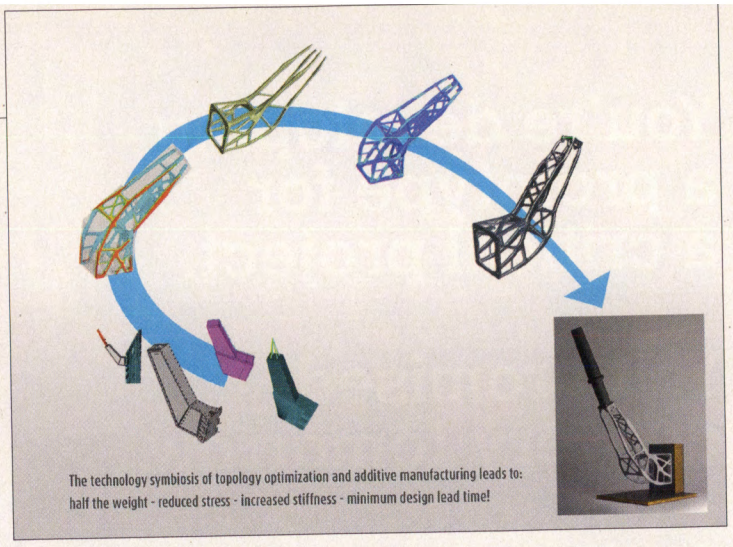
satellites and that its development isn't a "one-off" idea. Further space applications for industrial 3D printed components could include entire satellite structures—meaning that components like electrical harnesses, reflectors, and heating pipes (which still must be manufactured individually) could be integrated into standard 3D-printed structural elements.



To construct the metal 3D print, built-up layers of powder were joined together to form the desired shape via automated laser sintering. *(Image courtesy of Altair)*

3D RADAR'S LONG-RANGE HAZARD DETECTION Quickens Flight Rerouting

While traditional radar systems typically focus on one specific area at a time, a new 3D weather radar can analyze data from 17 different angles simultaneously. Honeywell's IntuVue 3D radar utilizes volumetric 3D scanning and pulse-compression technologies to provide pilots with a more complete view of the weather, at a greater resolution. This helps improve strategic maneuvering for rerouting and reduces the risk of encountering hazardous weather.

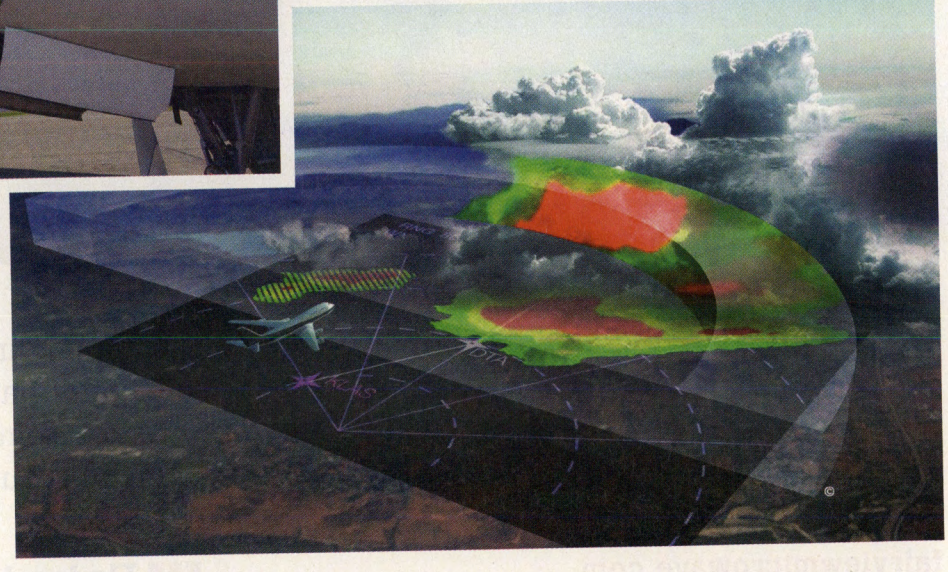
The IntuVue system, which fits into the nose cone of a plane, collects and automatically corrects data for earth curvature, offering a 180° display at 320 nautical miles from the aircraft. Automatic scan and ground clutter extraction helps ensure that the most reflective part of a storm is displayed, and that hazardous cells below the aircraft's flight level can be identified earlier and displayed longer than traditional systems. The system also includes advanced predictive windshear detection and alarming, which may reduce false alarms by about 90%. Its fully automated weather detection ultimately will reduce pilot workloads.

The aggressive down-tilt and 3D scanning process builds in inherent over-flight protection. IntuVue is also certified to the FAA Minimum Operating Performance Standard (MOPS) for enhanced turbulence—extended turbulence detection range reaches 60 nmi. Overall, the system helps to minimize the financial impact of severe storm activity while aiming to provide efficiency, safety, and performance. ■



The IntuVue system fits into the nose cone of a plane to analyze data from 17 different angles simultaneously. *(Images courtesy of Honeywell)*

The radar offers a 180° display at 320 nautical miles from the aircraft.



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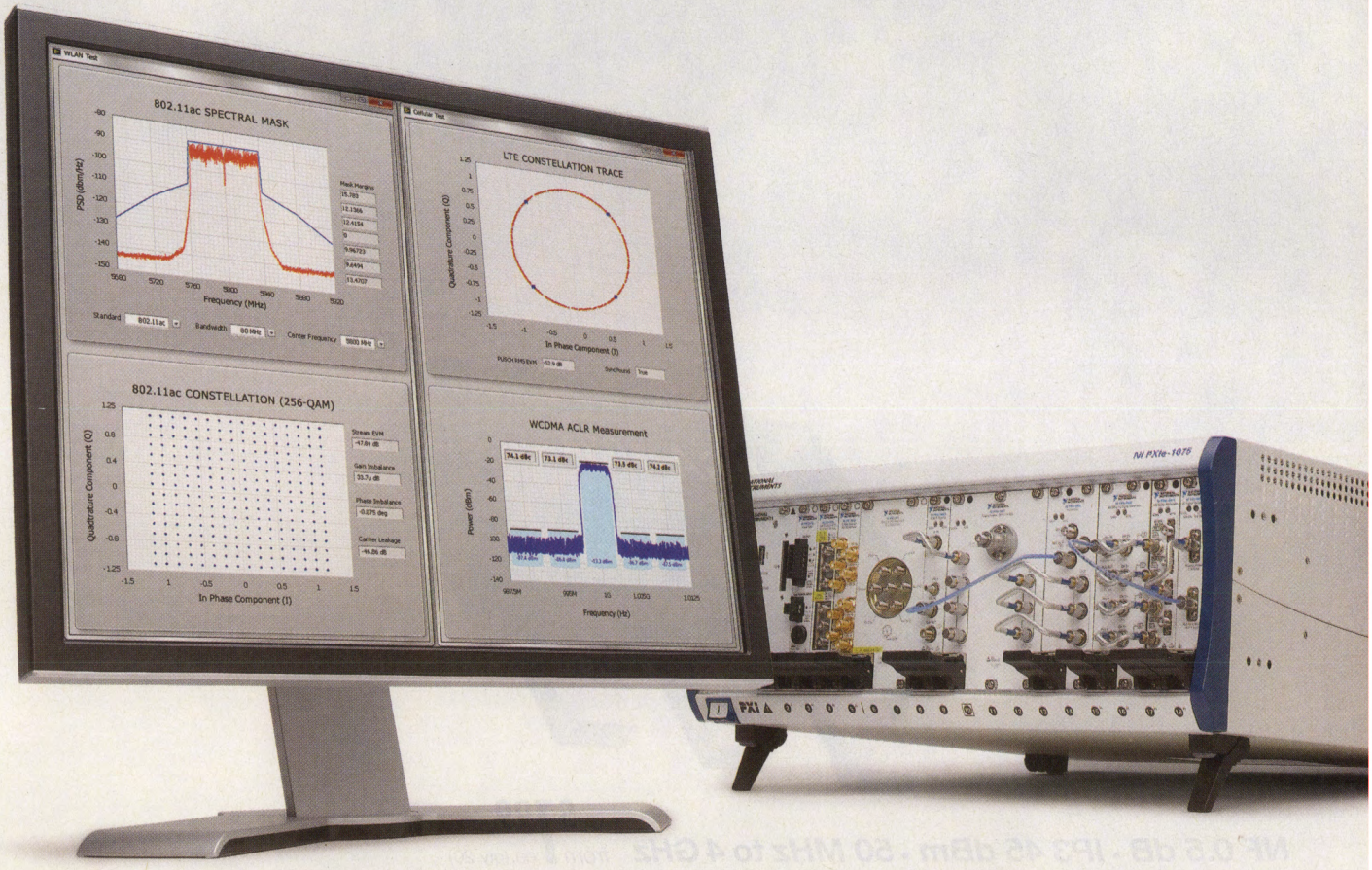

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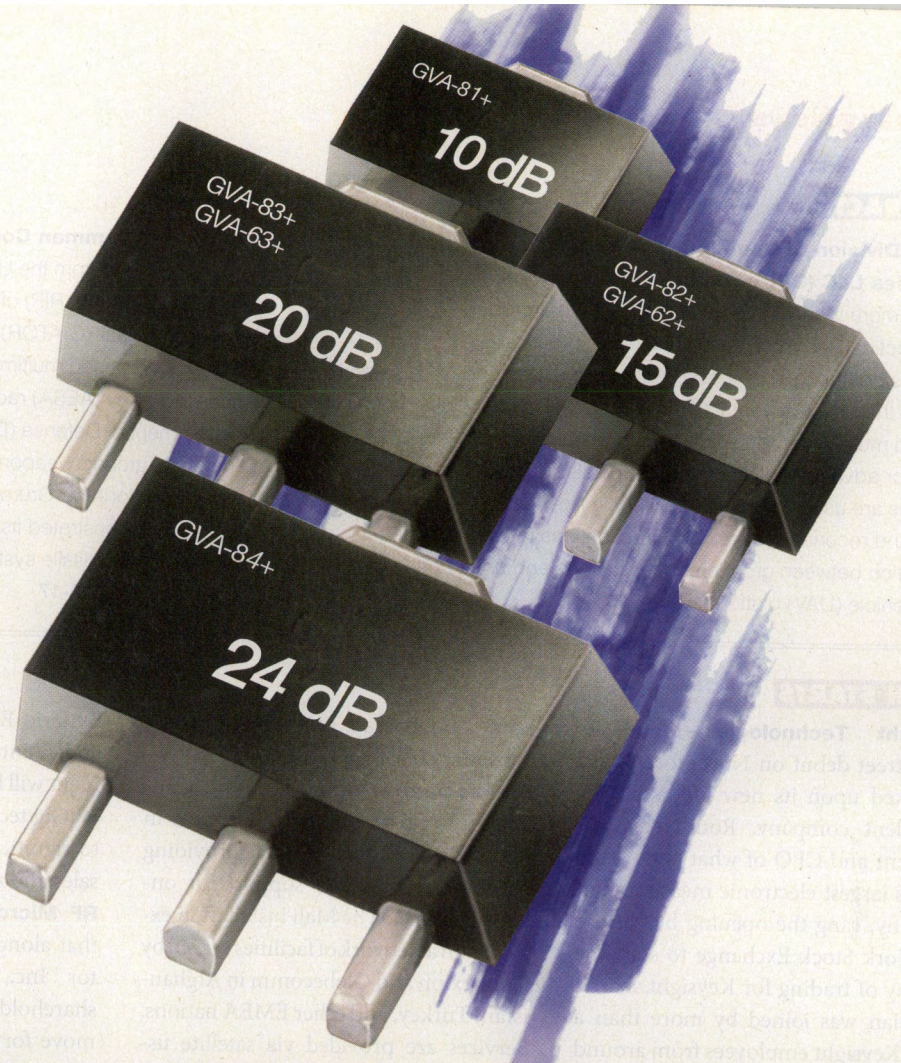

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CPI Malibu Reaches $30 Million Over Two Years

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Malibu Division of Communications & Power Industries LLC (CPI)—Was awarded a follow-on order of more than $5 million for the continued production of tactical common data link antenna ground terminals operating in Ku-band. With this most recent order, in the past two years, CPI Malibu Division has received multi-year orders totaling more than $30 million for advanced ground data terminals. These ground data terminals are used by the U.S. military in intelligence, surveillance, and reconnaissance (ISR) applications to communicate information between ground control stations and unmanned aerial vehicle (UAV) platforms.

Northrop Grumman Corp.—Received a $207-million contract from the U.S. Marine Corps for low-rate initial production (LRIP) of AN/TPS-80 Ground/Air Task Oriented Radar (G/ATOR) systems. G/ATOR is the first ground-based multimission active electronically scanned array (AESA) radar to be developed by the Department of Defense (DOD). It provides air surveillance, air defense, ground weapon locating and air traffic control capabilities. Under Northrop Grumman funding, G/ATOR has also demonstrated its ability to reliably detect high-speed rocket and missile systems. Northrop Grumman will deliver the systems in 2016-17.

FRESH STARTS

Keysight Technologies—Made its Wall Street debut on November 3 and embarked upon its new life as an independent company. Ron Nersesian, president and CEO of what is now the world's largest electronic measurement company, rang the opening bell at the New York Stock Exchange to start the first day of trading for Keysight shares. Nersesian was joined by more than a dozen Keysight employees from around the world who spent the past year working to separate the business from parent company Agilent Technologies.

Fairview Microwave Inc.—Launched its new RF/Microwave eCommerce website (www.fairviewmicrowave.com) featuring a powerful specification-based navigation, enhanced site-search capabilities, and a newly designed user interface.The website allows users to easily navigate through more than 30 RF product categories with its specification-rich parametric navigation. Product pages feature high-resolution photos, enhanced datasheets, applicable certificates, and real-time inventory levels and pricing.

SiTime Corp.—Completed its acquisition by MegaChips Corp., a fabless semiconductor company based in Japan. This closing follows the announcement on Oct. 28, 2014, that SiTime had entered into a definitive agreement to be acquired by MegaChips for $200 million in cash. SiTime will retain its name and operate as a wholly owned subsidiary of MegaChips.

Globecomm—Partnered with Gannexion to provide e-welfare services in the African nation of Mali, providing communications and support for on-site personnel. The Mali installation expands on a network of facilities served by Gannexion and Globecomm in Afghanistan, Turkey, and other EMEA nations. Services are provided via satellite using Globecomm's UK and Netherlands teleport, and each remote site features a satellite dish, Wi-Fi, and GSM antennas for local delivery of radio, television, voice, and data services, as well as video teleconferencing sessions (VTC).

Anritsu Co.—Expands its training courses with the introduction of Interference Analysis Certification, a training class to help wireless professionals locate interference sources. Attendees will learn about theory, procedure, calculations, instrument use, measurement parameters, trace interpretation, case studies, and interpersonal communication strategies associated with interference analysis. All students who pass their practical exam will be Anritsu certified, and receive a photo ID badge and certificate of course completion.

Peraso Technologies Inc.—Has closed a $20 million funding round led by new investor Roadmap Capital, with additional financing from existing investors Celtic House Venture Partners, Ontario Emerging Technologies Fund, and VentureLink Funds. The investment will be used to bring all of Peraso's WiGig technology into production and to significantly increase the company's sales and marketing activities.

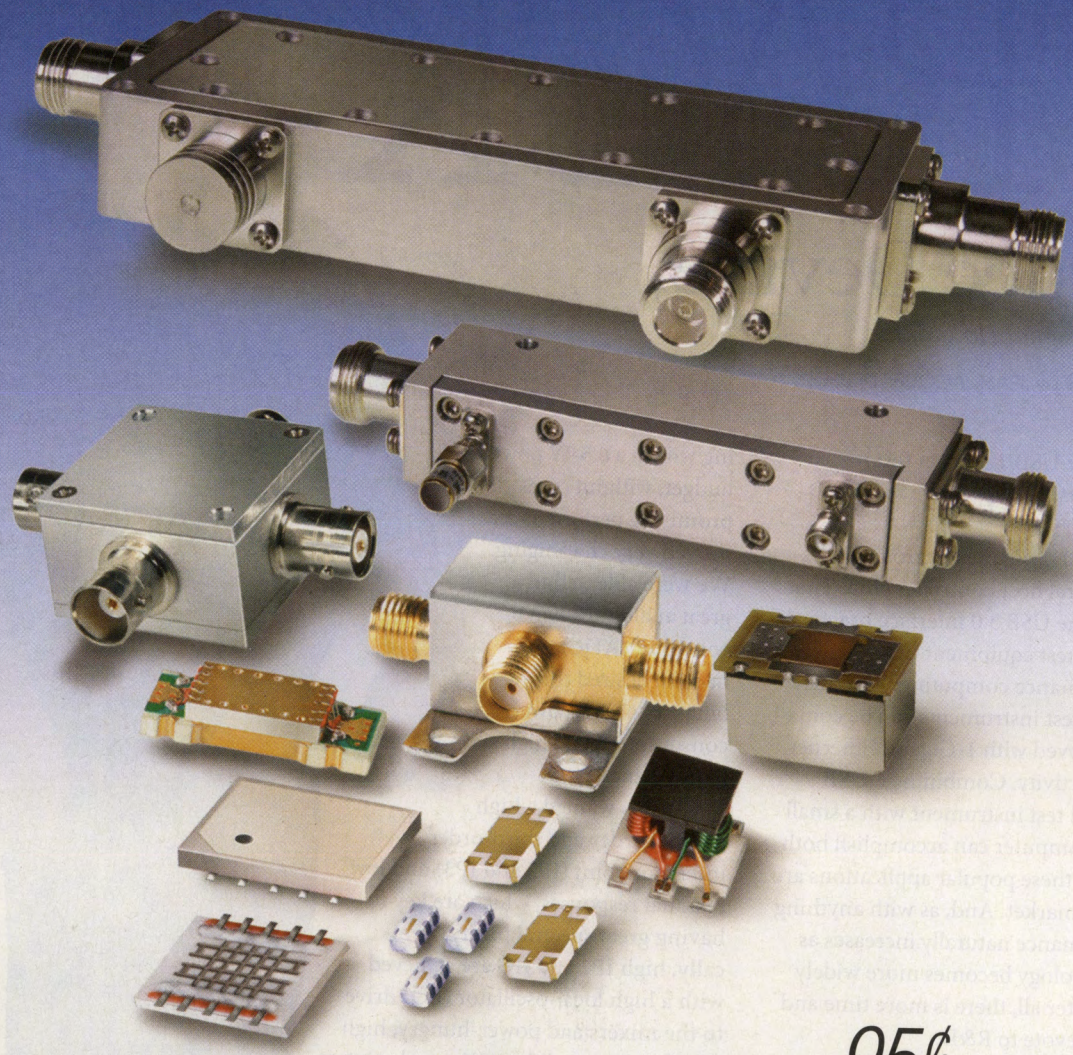
RF Micro Devices Inc.—Announced that along with TriQuint Semiconductor Inc., it has received all necessary shareholder and regulatory approvals to move forward with the previously announced merger of the companies and has set Dec. 31, 2014, as the anticipated closing date for the transaction.

Richardson RFPD Inc.—Introduced a new tuning kit for embedded GPS helix antennas from Maxtena Inc. The kit contains five standard HeliCore samples optimized for different loading conditions commonly encountered in devices requiring an embedded antenna configuration. It allows for quick and easy antenna selection and removes the need for a lengthy and costly custom antenna integration process.

STMicroelectronics—Has begun production of MEMS sensors using its proprietary THELMA60 (60µm Thick Epipoly Layer for Micro-gyroscopes and Accelerometers) surface-micromachining fabrication process. The ST innovation combines the advantages of the two most effective approaches, opening up the possibility to drive surface-micromachined MEMS into new markets and applications.

Directional / Bi-Directional

COUPLERS



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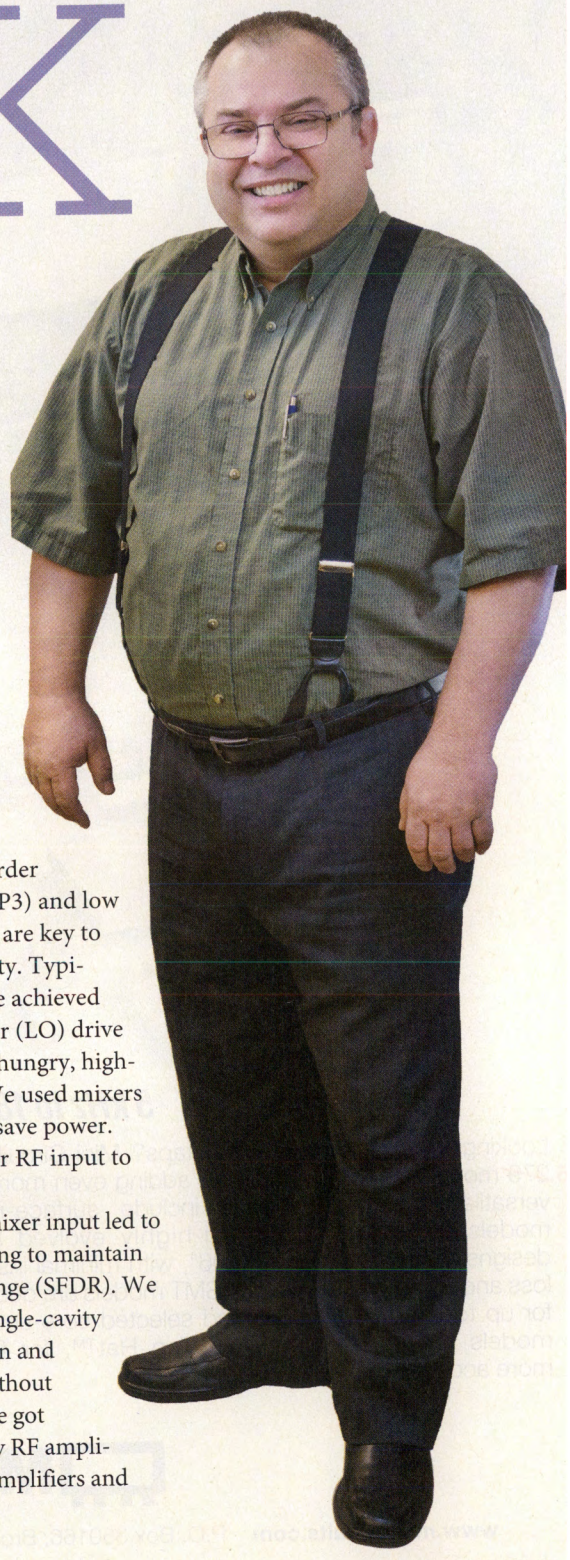
Inside TRACK with Bruce Devine,

CEO, Signal Hound, Inc.

Interview by JEAN-JACQUES DELISLE



JJD: Why are USB/Ethernet test instruments increasing in performance and availability?

BD: In late 2012, Cypress introduced its FX3 peripheral host controller chip. That chip made the USB 3.0 interface practical for use with test equipment that requires high-performance computing. Remotely controlling test instruments can be conveniently achieved with 1-Gigabit Ethernet (GE) connectivity. Combining a USB 3.0-powered test instrument with a small industrial computer can accomplish both tasks. All of these popular applications are driving the market. And, as with anything else, performance naturally increases as a new technology becomes more widely adopted. After all, there is more time and money to devote to R&D.

JJD: When designing a compact USB spectrum analyzer, what challenges must be overcome to maintain the power budget?

BD: In spectrum-analyzer design, a single USB 3.0 port is limited to 5 V and 900 mA for a total of 4.5 W. Clearly, the single-port power limit was inadequate for the high-performance design of our BB60C spectrum analyzer. A USB3.0 Y-cable was therefore required to provide more current. Our BB60C design draws 1.3 A or 6.5 W. That's as low as we could bring the power budget without losing valuable performance. Staying within a 6.5-W power budget, without compromising performance, proved very challenging. We were able to find a great analog-to-digital converter (ADC) that had the desired combination of low power consumption and high performance.

We also wanted a high second-order and third-order intercept point (IP2 and IP3) and low residual responses, which are key to having great spectral purity. Typically, high IP2 and IP3 are achieved with a high local-oscillator (LO) drive to the mixers and power-hungry, high-linearity RF amplifiers. We used mixers with a lower LO drive to save power. That required much lower RF input to keep down the spurious.

The lower-power RF-mixer input led to the need for better shielding to maintain spurious-free dynamic range (SFDR). We used several one-piece, single-cavity shields to provide isolation and maintain performance without adding too much cost. We got around the power-hungry RF amplifiers by using push-pull amplifiers and

adding more RF leveling control. This allowed us to keep the mixer RF-input levels in a range where second- and third-order products are at least 50 dB down over a wide range of input levels.

We also had to model distributed-element bandpass and lowpass filters, which we embedded into the printed-circuit-board (PCB) laminate inner layers to save shielding space and still benefit from their filtering. In the end, filtering and shielding were more challenging than fighting the power budget because of the added constraints of size and cost. Designing for minimal size, cost, and power while maintaining high performance is not a trivial endeavor.

JJD: How have customers been embracing these new devices' compact footprint and rich features?

BD: Although size is an important feature to many customers, price appears to trump everything. Customers were willing to embrace USB-powered test equipment initially because the price was right. They became used to seeing other companies' USB-powered spectrum analyzers, which had questionable performance and weren't taken very seriously. Then, in February 2010, we introduced the USB-SA44 spectrum analyzer.

People didn't know what to think about the USB-SA44. Our price got them in the door, so they were brave enough to try it out. Upon using it, they were delighted to discover that it did what they needed and performed as advertised. I have taken numerous calls over the years in which someone asked me if our products really did what we said they could do. I finally had to resort to offering a 30-day money-back guarantee just to strengthen our credibility.

Now, five years later, enough momentum has gathered that the notion of serious USB-powered test equipment is no longer foreign. More customers are embracing USB-powered test equipment. With its compact footprint and rich features, it offers something that was unaffordable when one-box solutions were the only realistic choice. USB-powered test equipment is gaining respect because the customer finally has a full range of serious instruments available.

JD: What unique new markets have been enabled by these devices?

BD: Universities are now able to afford to stock their labs with real-time spectrum analyzers. This provides engineering students with access to both a curriculum and hands-on experience that were previously prohibitively expensive. In addition, automated test equipment (ATE) has become more affordable for manufacturers to use on their production lines.

"Designing for minimal size, cost, and power while maintaining high performance is not a trivial endeavor."

One of our customers manufactures small wireless devices. They have designed-in more than 50 of our USB-SA44B spectrum analyzers for use in production-line testing. In many instances like theirs, a traditional spectrum analyzer is overkill and a waste of money. A USB-powered real-time spectrum analyzer's persistence function is perfect for troubleshooting and finding intermittent signals.

JJD: With many of these devices within a hobbyist's budget, what interest have they garnered in the hobby corner?

BD: Ham-radio operators favor our USB-SA44B/USB-TG44A combination of spectrum analyzer and tracking generator. It is economical and the pair works together to form a scalar network analyzer. It allows them to plot filters, calibrate step attenuators, perform AM/FM modulation measurements, and—with the addition of a directional coupler—make antenna and cable insertion-loss and return-loss measurements.

JJD: Beyond USB2.0/USB3.0/Ethernet, what interface technologies are being reviewed for use? And what factors need to be considered regarding interface technology?

BD: These interfaces provide enough data throughput for all but the most demanding applications. Real-time spectrum analyzers are using USB 3.0 because it has the highest throughput of the commonly available interfaces mentioned. There is nothing wrong with directly connecting to PCI Express for desktop PCs, but it is difficult if not completely impractical for use with laptops. Computer manufacturers will begin integrating 10 Gigabit Ethernet into their laptop peripheral controllers, so it becomes as commonly used as USB3.0. At that point, we will consider using it as the next stable and practical interface for test-equipment architectures that require high-performance computing.

U.S. export controls will also have to be revised, as test equipment using 10 Gigabit Ethernet is currently restricted for export due to the technology being too powerful. Once U.S. commerce controls allow the unrestricted export of test equipment with a 10 Gigabit Ethernet connection, you will start to see affordable real-time spectrum analyzers with 100 MHz of instantaneous streaming bandwidth. But for now, USB 3.0 is king.

JJD: What opportunities can we expect with PC-driven test instruments?

BD: Portability. For example, teaming a USB 2.0-equipped quad-core tablet PC with a USB test instrument produces a very effective field instrument. We're already doing it at our factory. Once we release our rewritten SA/TG series software this January, customers will be able to make the same pairing. The SA/TG rewrite will be closely followed by the introduction of a touchscreen user interface. Users will be able to Velcro the USB-SA44B to the back of a tablet PC and have a spectrum analyzer for field work. The same possibilities exist for other manufacturers that offer USB-powered test equipment, such as oscilloscopes and signal generators. MW

MIMO RADAR COULD OUTPERFORM PHASED-ARRAY RADAR

ANOTHER STEP IN the evolution of radar may be the use of multiple-input multiple-output (MIMO) technology, which utilizes spatially diverse transmitters and receivers. Mark T. Frankford, Kyle B. Stewart, Ninoslav Majurec, and Joel T. Johnson, with support from the Swedish Research Council and the Swedish Foundation for Strategic Research, recently investigated the use of MIMO radars. They found that these radars may offer enhanced performance with better tolerance to fading, as the target's radar cross section (RCS) at the aspect angle is varied.

Through both simulation and experimentation, the researchers investigated whether MIMO radar systems could outperform phased-array radar systems in higher signal-to-noise-ratio (SNR) environments. Their simulation study used a FEKO computational electromagnetics package on a helicopter-like target at 200 MHz with two transmitters and four receivers. For the experimental study, an unmanned-aerial-vehicle (UAV) target was measured using a software-defined-radio (SDR) platform at 2.75 and 4.50 GHz.

For these two tests, the study confirmed that a MIMO radar approach performs better at high SNR values using spatially diverse antennas. Further investigation will explore the extension of Ohio State's SDR technology to four transmit channels and four instantaneous receive channels for a deeper understanding of MIMO radar capabilities. See "Numerical and experimental studies of target detection with MIMO radar," *IEEE Transaction on Aerospace and Electronic Systems*, April 2014, pg. 1569.

DE-EMBED PARASITICS FROM MILLIMETER-WAVE ICs

WHEN CHARACTERIZING AN integrated circuit, it is critical to de-embed parasitics and test-structure features. A team from the University of Grenoble Alpes in France—Vipin Velayudhan, Emmanuel Pistono, and Jean-Daniel Arnould—has proposed a highly accurate de-embedding method based on a half-through approach.

Their method models the probe pads and interconnects using a matrix calculation, which is obtained directly from the S-parameters as test fixtures are de-embedded. This method removes the need to assume symmetry and equivalent models for pad-interconnect parasitics. To create the de-embedding model, three different S-parameter tests must be performed. First, a transmission line with ground-signal-ground (GSG) pads that are 100 µm in length must be tested for transmission and reflection. The same testing is then performed on a GSG transmission line that is 200 µm in length. Finally, a reflection test is performed on a half-through line that is terminated in a non-characteristic load.

The paper discusses the derivation of the equivalent model by converting S-parameters to ABCD matrices. Using the reflection coefficient of the half-through test and signal-flow graph theory, or Mason's Rule, the researchers then extract the effects of the pads and interconnects referenced to 50 Ω. The paper claims that the half-through method outperforms other methods under simulation. The remaining work is to physically implement the half-through method to confirm performance expectations.

See "Half-Thru de-embedding method for millimeter-wave and sub-millimeter-wave integrated circuits," *2014 10th Conference on Ph.D. Research in Microelectronics and Electronics (PRIME)*, June 2014, pg. 1.

INP RESONANT TUNNELING DIODES DROP LNA NOISE FIGURE

ONE OF THE challenges for devices with extremely low power and adequate RF performance is the need to maintain low power while operating the necessary low-noise amplifiers (LNAs) for signal reception. Using resonant tunneling diodes (RTDs), a negative-differential-conductance (NDC) amplifier has been designed by Jongqon Lee, Jooseok Lee, Maengkyu Kim, and Kyounghoon Yang from the Department of Electrical Engineering of KAIST in the Republic of Korea.

Indium-phosphide (InP) RTDs possess good NDC characteristics at low DC power. This feature makes them suitable for microwave wireless transmission for remote and sensor applications, which require high performance in power-restricted applications. The research team uses two RTD amplifiers joined with a hybrid coupler. The bypass capacitors, thin-film resistors, spiral inductors, and RTDs in the device are all integrated with a 0.9-µm InP monolithic-microwave-integrated-circuit (MMIC) process.

The center frequency for the low-power LNA is centered at 5.5 GHz with gain of 8.1 dB, a noise figure of 4.5 dB, and a gain-to-DC power ratio of 52.2 dB/mW. These performance metrics exceed comparable LNAs. The next closest 180-nm CMOS process reportedly reaches just a 12.8 dB/mW gain-to-DC power ratio. See "Negative-differential-conductance RTD amplifier MMIC with record forms of gain-to-dc power ratio and noise figure," *26th Conference on Indium Phosphide and Related Materials (IPRM)*, May 2014, pg. 1.

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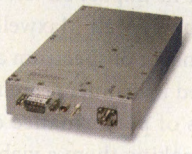

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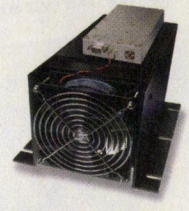

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Special Report

JEAN-JACQUES DELISLE | Technical Editor

PC-Driven Test PICK UP SPEED

The trend toward more integrated test instruments has improved performance at the high end while spawning a market for low-cost, portable, and PC-driven instruments.

Bucking the trend of rising costs, clever test-and-measurement designers have been successfully exploring methods of reducing costs and still providing quality performance instruments. For example, a new market for test instruments is being created by the separation of an instrument's processing/control and measurement modules. With the accelerated performance of personal computers, it also has proven viable to create PC-driven instruments that interface through high-speed and common platform-interface technologies.

Until recently, there were several technological barriers to using PC-driven techniques for high-performance test-and-measurement instruments. One such restriction was the inability to rapidly transfer data through a common interface platform from a sensor to a PC. Here, speeds had to be fast enough to provide adequate fidelity. Additionally, CPU processing speeds, physical memory, and random-access memory (RAM) were not up to the speed or capacity challenge of handling professional-grade instrument data output. These factors would lead to a large bottleneck as attempts were made to handle the data from the instruments. In addition, limitations were placed on the processing of data once it was transferred. Another challenge associated with portable platforms was providing enough power for both the test instrument and PC hardware. In the mid-2000s, a few components could use the power and data rates offered by the technology, such as those based on USB 2.0. Yet network analyzers (NAs) or spectrum analyzers (SAs) with enough bandwidth or frequency range to compete with even low-end benchtop instruments were still inconceivable.

To solve many of these limitations with current hardware would require a proprietary interface technology and a specialized workhorse of a computer, mitigating the benefits of a PC-driven computer. In 2008, the USB-IF released USB 3.0, an update to the USB standard that enabled raw data rates of 5.0 Gbps and 900 mA of simultaneous power draw. The PC world adopted USB 3.0 as the standard for the latest computers, laptops, and even several tablet computers. As for its impact on test and measurement equipment, Matt Maxwell, Tektronix's product manager of spectrum analyzers, notes, "USB-based instruments are no longer limited in terms of performance or analysis capabilities, making them a viable option for much more demanding applications than in the past."



1. In what used to be the size of a standard Flash drive, some compact spectrum-analyzer manufacturers are enabling spectrum sensing to 8.15 GHz. *(Courtesy of Triarchy)*

The low costs and high data capabilities

Instruments

associated with USB have supported the emergence of a variety of different applications. Examples range from USB test instruments to high-definition/high-speed imaging cameras and industrial sensing/control. Maxwell says, "Until the advent of USB 3.0, it simply wasn't possible to move real-time signal information to a laptop or tablet computer fast enough. With highly capable quad-core processors like the Intel Core i7—along with solid-state drives (SSDs) and fast RAM readily available at affordable price points—today's USB 3.0-equipped PCs and tablets represent capable platforms for supporting test-and-measurement applications."

Portable and low-power processing, memory, and storage have all advanced, thanks to the growing demand for portable computers and the Internet-of-Things (IoT) connected environment. Meanwhile, higher-performance and smaller, solid-state memories and CMOS microelectronics have reduced both their size and power draw on computers. In fact, many of the latest laptops/tablets have nearly desktop-equivalent processing capability. The portability of heavy processing has enabled convenient on-site programing, processing, and data storage for large amounts of complex sensor data.

As computers became a standby in every household and office, the number of computer accessories and peripherals also grew. The size of these markets drove computer manufacturers to settle mostly on common interfacing, which has enabled PC-driven test and measurement instruments to advance more consistently. Maxwell says, "Test-and-measurement providers are experts at designing and building the actual signal-acquisition and analysis solution. PC manufacturers are experts at designing and building computer systems, and can do so more cost-effectively than test-and-measurement providers."

Take the use of commercial-off-the-shelf technology, which is increasingly being used by industrial and military organizations. With computers as well as other consumer products, modular components have decreased in cost and increased in performance at a rapid pace that proprietary technologies could not maintain. "Therefore, it makes economic sense for test-and-measurement companies to take advantage of the advances enabled by the PC industry," says Maxwell. "By removing the requirement to build a front-end PC, test-and-measurement companies can offer instrumentation at lower price points and bring new products to market faster." (*Fig. 1.*)

The benefits of PC-driven test-and-measurement instruments, such as cost, portability, and rich software potential, are attractive enough to encourage early technology innovators. Yet new techniques must be used to shrink the footprint of such high-performance electronics with a limited power budget and the necessary ruggedization. "USB instrumentation needs to be light, compact, and rugged," explains Maxwell. "This typically requires all new board layouts and potentially new ASICs, along with close attention to heat management while ensuring a clean signal path. While USB 3.0 is fast, bandwidth limitations and power constraints can limit performance."

Even with USB 3.0's increased data rate, practical limitations prevent the interface from reaching the maximum 5-Gbps data rate. Practical data rates range to 3.2 Gbps (*Fig. 2*). Yet Maxwell says, "The bandwidth limitations can be overcome by using a faster interface than USB 3.0, such as USB 3.1, Thunderbolt, or even multiple connections."

SOFTWARE ADVANCES

To ensure the adoption of PC-driven instrumentation, it is critical that a human interface allows an engineer or technician to quickly become familiarized with the instrument



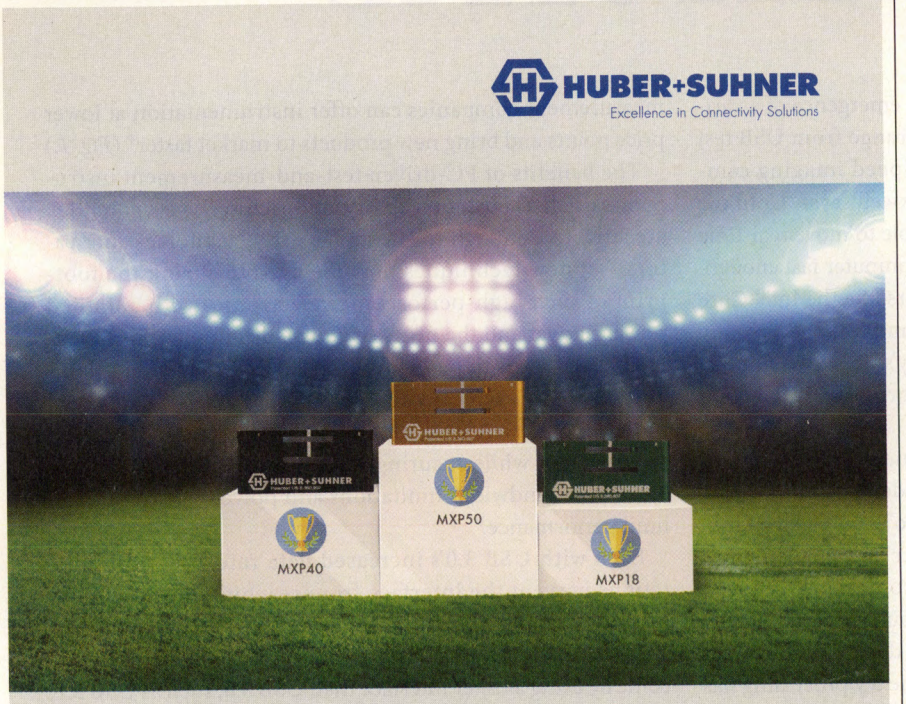
2. By offering modules for either USB or Ethernet-based connectivity, some PC-driven test and measurement manufacturers are taking a more modular approach to increase data rates for a variety of instruments. *(Courtesy of Holzworth Instrumentation)*

controls (*Fig. 3*). Of course, software-based controls offer great potential in terms of adding more customizability, features, and performance than a physical interface. The user, however, ultimately decides how an interface must be designed. Maxwell emphasizes, "Advanced capabilities in software are vital to the success of USB instrumentation. Without a front-panel display, software for controlling the instrument must be fast, intuitive, and easy to use. Usability studies show that productivity is similar across instruments that include hardware controls and software-only based instruments."

An added benefit of this class of instruments is that the software behind the instrument display is often subject to manipulation and even outright redesign by the user. Access to software development kits (SDKs) is occasionally provided by manufacturers of PC-driven instruments. It enables users to completely manage the processing and display of the data fed by the instruments. For large-scale manufacturing environments, this flexibility could be used to develop rapid or even automated testing processes that demand much less development investment than benchtop instruments.

Examples of USB 3.0-connected instruments include new portable and real-time SAs from Signal Hound and Tektronix. Signal Hound's BB60C SA streams 140 MB/s of digitized RF spectrum from 9 kHz to 6 GHz, enabling an instantaneous bandwidth of 27 MHz at sweep speeds of 24 GHz/s. The Tektronix RSA306 SA captures signals from 9 kHz to 6.2 GHz with 40 MHz of real-time bandwidth down to −160 dBm. Both instruments come with a built-in software package and application programming interface (API) for customizable software and interfacing. The RSA306 also includes a Matlab driver for the API, which can readily connect with Matlab's Instrument Control Toolbox.

Both of these devices are recommended for use with PCs that have an Intel i7 processor and 8 GBs or more or RAM, which are needed to process the data. Because a main design consideration of these devices is to lower end-user costs, USB 3.0 is solely used. Yet the addition of a compact computer with a network interface can transform these instruments into full-functioning portable test instruments with remotely programmable and controllable capabilities. In addition, Gigabit USB device servers enable 10/100/1000 Mb of switched Ethernet connectivity with higher data rates eyed for the future.



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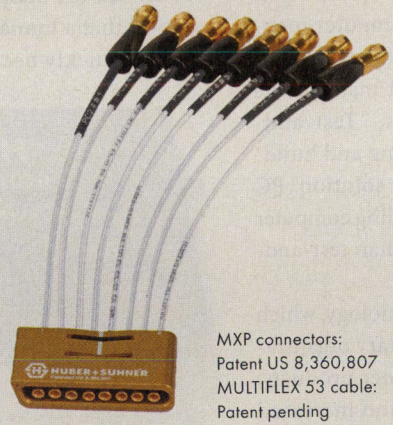

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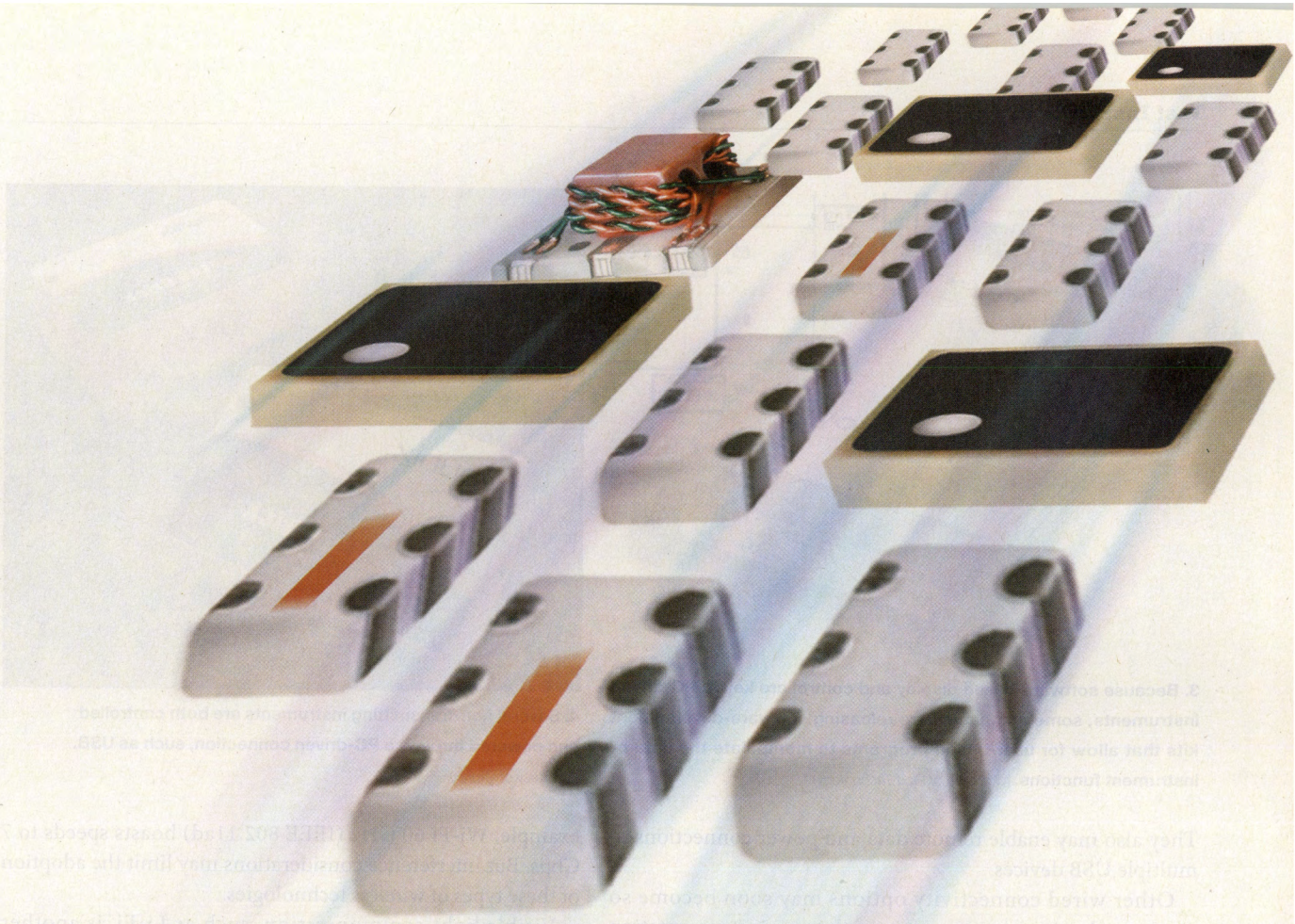

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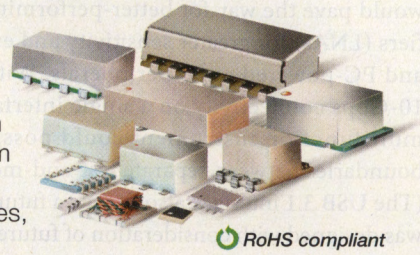

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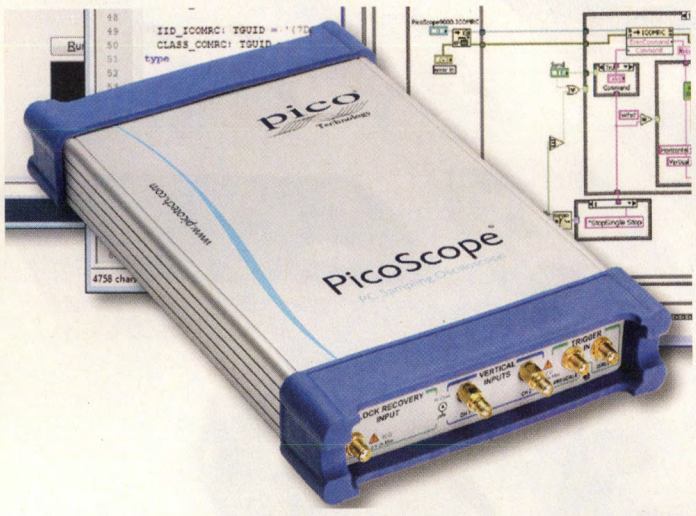

3. Because software-based display and control are key to PC-driven instruments, some companies are releasing software-development kits that allow for user-based programs to manipulate the data or instrument functions. *(Courtesy of Pico Technology)*



4. Several non-transmitting instruments are both controlled and powered through a PC-driven connection, such as USB. *(Courtesy of Vaunix)*

They also may enable remote data and power connections to multiple USB devices.

Other wired connectivity options may soon become so common that it makes sense to use them as instrumentation interfaces. The most likely to impact PC-driven test equipment is USB 3.1, which increases USB 3 data rates to 10 Gbps and reduces the line-encoding overhead to 3%. That reduction in overhead results from the change to the 128b/132b encoding scheme. Although the data rate is crucial, the new USB 3.1 standard also will allow for multiple power levels: 10 W with 2 A at 5 V, 60 W with 5 A at 12 V, and 100 W with 5 A at 20 V.

This 2× to 20× increase in power would enable much more power-hungry and higher-speed digital-to-analog converters (DACs) and analog-to-digital converters (ADCs). It also would pave the way for better-performing low-noise amplifiers (LNAs) for greater sensitivity and even fully PC-driven and PC-powered radio peripherals. With up to 100 W and 10-Gbps connectivity on a single interface, a cost-effective multifunction instrument could possibly tear down the boundaries between separate test-and-measurement devices. (The USB 3.1 interface also flaunts a future-proof design, as it was designed with consideration of future speeds greater than 20 Gbps, according to USB-IF.)

Two main bottlenecks remain for PC-driven instrumentation: data rates and the adoption of interfaces that can transfer the data needed to enable high-performance instruments. Another path may be to move toward wireless interface standards, as the consumer market is demanding ever-higher bandwidths from wireless connections. Indeed, there have been significant increases in data rates and bit-error-rate (BER) improvements with wireless interface technologies. For example, Wi-Fi 60 GHz (IEEE 802.11ad) boasts speeds to 7 Gbps. But interference considerations may limit the adoption of these types of wireless technologies.

Visible light communication, such as Li-Fi, is another emerging technology that could be free of electromagnetic interference. This technology uses visible light-emitting diodes (LEDs) to transmit and receive data. Current experimental rates range to 10 Gbps. Much faster speeds are theorized, due to the huge bandwidth availability and low interference in visible bands.

LOOKING FORWARD

Buoyed by new developments, PC-driven test-and-measurement instruments will continue to overcome any doubt and hesitancy in the engineering community. Already, most major test-and-measurement companies have embraced PC-driven architectures for low- and mid-range network analyzers, spectrum analyzers, and other components (*Fig. 4*). "Based on these advantages, we're confident that the trend toward professional-grade USB instrumentation will continue and have a significant impact on the market," says Maxwell. "As faster interfaces, such as USB 3.1, become mainstream, look for more capable and higher-performance instruments to be introduced. Meanwhile, demand for remotely connected/controlled devices is increasing in the industrial and commercial environments, which may lead to more devices being equipped with either wireless or wired Ethernet capability.

"Ethernet connectivity may also be a possibility if there is shown to be sufficient need and demand," Maxwell continues. "To be sure, the future is indeed bright for USB/Ethernet instrumentation." MW



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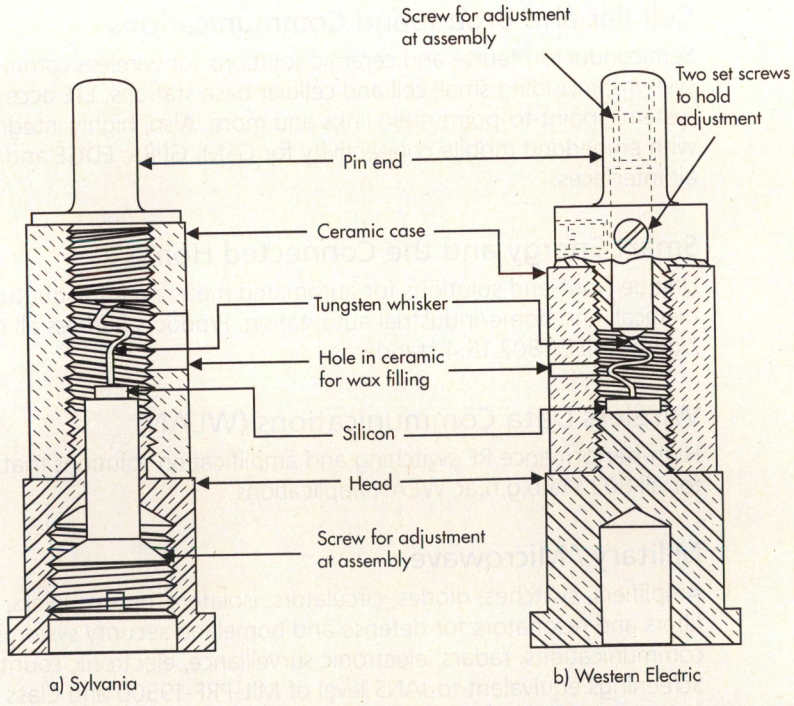
JEAN-JACQUES DELISLE | Technical Engineering Editor

A Closer Look at RF Power Measurements

Because measuring RF/microwave power is one of the most commonly used methods of understanding system performance, detailed knowledge of how power measurements are made is key to successfully interpreting measurement results.

POWER IS IMPORTANT. Measuring the power from an RF/microwave signal is a fundamental way to gain knowledge about signals and devices. Understandably, it is critical to accurately and consistently measure power. It also is essential to maintain a standard of power measurement that enables measured power to be communicated to customers, manufacturers, designers, and researchers. There are three main technologies involved in measuring RF/microwave power: the diode, thermistor, and thermocouple sensor. Each product type comes with its own benefits and uncertainties. Additionally, industry and manufacturer standards for power-measurement calibration need to be employed successfully to communicate absolute values of power (namely, the National Institute of Standards and Technology and other National Measurement Institutes). Having a clear understanding of the fundamentals of power measurement, including the technology and standards, will ensure that an engineer has the most reliable information for optimizing their designs.

The goal of a power sensor, for example, is to absorb the incoming energy from an RF/microwave signal. By relying on energy conversion, it then attempts to transfer that energy into usable electrical energy that exactly replicates the absorbed energy. Of course, some energy is always lost in the impedance mismatch between the source and sensor. In addition, some energy is absorbed in parts of the sensor that do not render usable electrical energy. The meter itself also suffers from some conversion inefficiencies. Fortunately, however, there are calibrations provided by power-meter manufacturers that use a chart or built-in memory to help compensate for the sensor and meter inefficiencies.



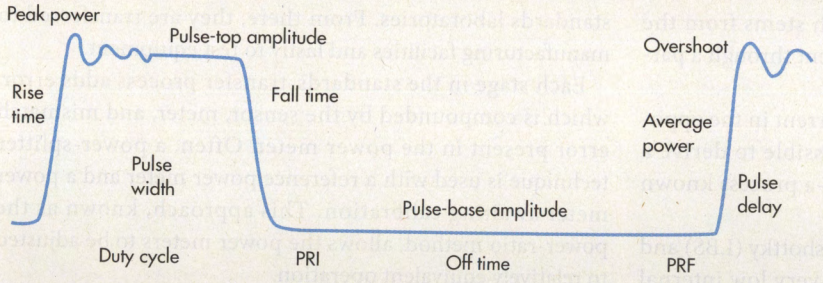
1. This microwave-sensor approach used silicon crystals with a curled tungsten wire, which touched the surface of the silicon functioning as a diode.

> "To achieve the lowest error possible for any of the measurement technologies, it is critical to account for the error in each stage of the sensor and meter process."

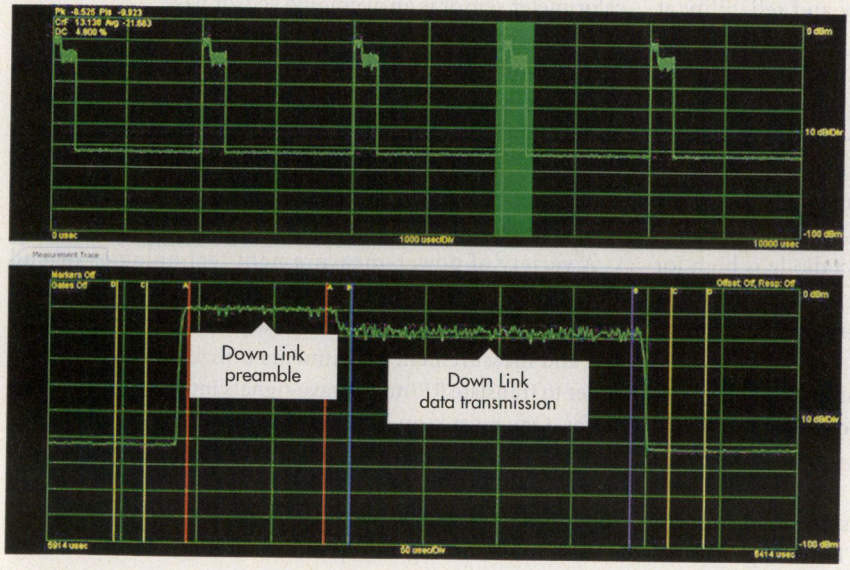
AN OLD-TIMEY THERMISTOR, STILL IN DEMAND

Of the three types of sensor technologies, the thermistor—a type of bolometer—has been around the longest. It is still used in many applications including science and research. The thermistor sensor converts RF/microwave energy received by a semiconductor-based material, which changes resistance as a function of temperature *(Fig. 1)*.

In addition, a feedback loop is used with a balanced-bridge typology. That typology employs a direct-current (DC) or alternating-current (AC) bias supply as a control to maintain a constant resistance for the sensor. As a result, the changes in bias power can be used as an indicator for the energy absorbed in the sensor.



2. The response characteristics of a pulsed signal are detailed in order to properly describe the signal behavior for interpretation. *(Courtesy of Keysight Technologies)*



3. Modern power meters can intelligently trigger on wireless protocols to properly assess wireless-device operational performance. *(Courtesy of LadyBug Tech)*

LOWER ERROR WITH THERMOCOUPLE SENSORS

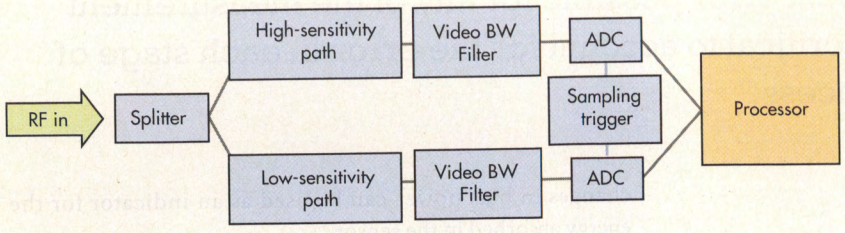
In many of today's applications, thermocouple sensors have replaced thermistors. This trend is most likely due to the advantages offered by thermocouple sensors: higher dynamic range, sensitivity, increased durability, and lower measurement uncertainties associated with standing wave ratio (SWR). Thermocouple operation relies on two physical bases—the first of which uses the voltage generated from the Peltier effect at the junction of two electrically dissimilar materials.

The other base leverages the Thomson effect at the junction of two different temperature sections of a conductor. Currently, metallic thin-film and semiconductor technology are used to create high-performing and linear thermocouples in very small footprints.

A transmission method is often employed to enable use of the extremely low voltage—sometimes as low as 1 nV/mW—generated by the thermocouple. One method is to chop the DC signal into a series of square waves and transmit that over an AC-coupled path after amplification *(Fig. 2)*.

Because this is an open-loop approach, an outside calibration method must be used. Otherwise, various aspects of the thermocouple sensor can lead to drift in the DC response to RF/microwave power.

Often, a highly accurate reference oscillator set to a specific known power is used between tests to calibrate the thermocouple-based sensor to match the reference source. Errors in the reference source, calibration mechanisms, or interconnect of the source and sensor can lead to measurement errors. It is therefore critical for the reference to



4. The modulation schemes common to the wireless industry may have extremely high peak-to-average power ratios. As a result, multipath diode sensors with high and low sensitivity paths can be used to measure average power. *(Courtesy of Tektronix)*

have as little error as possible as well as a verifiable trace history to a credible NMI.

DIODE DETECTION GROWS IN POPULARITY

Another power-measuring technology that is becoming more viable is the diode detector. The nonlinear current and voltage relationship within a diode, which stems from the junction properties, allows the flow of current through a particular path of the device.

At the same time, it resists the flow of current in the opposite direction. This behavior makes it possible to derive a positive waveform using multiple diodes—a process known as rectification.

Diode technologies, such as low-barrier-shottky (LBS) and planar-doped-barrier (PDB) diodes, have very low internal capacitance and low forward voltage. As a result, they can respond well to very rapidly changing signals *(Fig. 3)*. With advanced micro-fabrication techniques, these devices can be produced with a high level of consistency and reliability at relatively low costs.

Although many diode detectors are limited in the higher power regimes—beyond the −20 dBm level in which thermocouple sensors can accurately operate—diode sensors are often capable of accurate sensing down to −70 dBm. Thus, they are often chosen for high-sensitivity and high-frequency power measurements.

The resistance at the terminal of an LBS diode is high for low-signal levels and varies significantly with temperature. Consequently, additional matching circuitry may be required to achieve reliable results.

PDB diodes generally have consistent and much lower resistances than LBS diodes. They also boast lower internal capacitance, which leads to more stable square-law region behavior and a higher operating range.

With enhanced fabrication techniques, some diode sensors have reached 90 dB in dynamic range. This range is suitable for continuous-wave (CW) sensing into the millimeter-wave frequencies. Additional circuitry can be added to the diode power sensor on a monolithic-microwave-integrated-circuit (MMIC) substrate, which will allow advanced corrections to be applied to further increase the diode-sensor device performance.

To enable averaging power measurements, it is possible to use a high-sensitivity and low-sensitivity path connected to a splitter and then to the source *(Fig. 4)*.

ERROR & UNCERTAINTY

To maintain a common ground for power-measurement standards, NMI organizations have prescribed methods of obtaining calibration-standard traceability. These standards often start with extremely accurate micro-calorimeter standards. They are then transferred to working standards and on to measurement reference standards used by commercial standards laboratories. From there, they are transferred to manufacturing facilities and lastly to test equipment.

Each stage in the standards-transfer process adds error, which is compounded by the sensor, meter, and mismatch error present in the power meter. Often, a power-splitter technique is used with a reference power meter and a power meter awaiting calibration. This approach, known as the power-ratio method, allows the power meters to be adjusted to relatively equivalent operation.

Yet the asymmetries in the splitter calibration often add more error to the meter. More complex methods are needed to appropriately calibrate a peak-power sensor, as the micro-calorimeter sensor is an averaging sensor base.

To achieve the lowest error possible for any of the measurement technologies, it is critical to account for the error in each stage of the sensor and meter process. Precision assemblies can help to reduce the impedance mismatch between the sensor and source. In addition, multiple sensors can sometimes be used to mitigate the effects of bias, ambient temperature, and noise factors.

A variety of power-measurement technologies extend from bench-top to rack-mount power-measurement devices. Network analyzers, spectrum analyzers, and many other test and measurement instruments utilize power sensors in order to translate RF/microwave signals into human-relatable terms. One must keep in mind that the different techniques, technologies, and methods of power sensing may be better suited to measuring different power levels, frequencies, and modulations of signals.

As a result, having a deep understanding of the exact signals under scrutiny and available measurement techniques could aid in delivering an adequate quality result. mw

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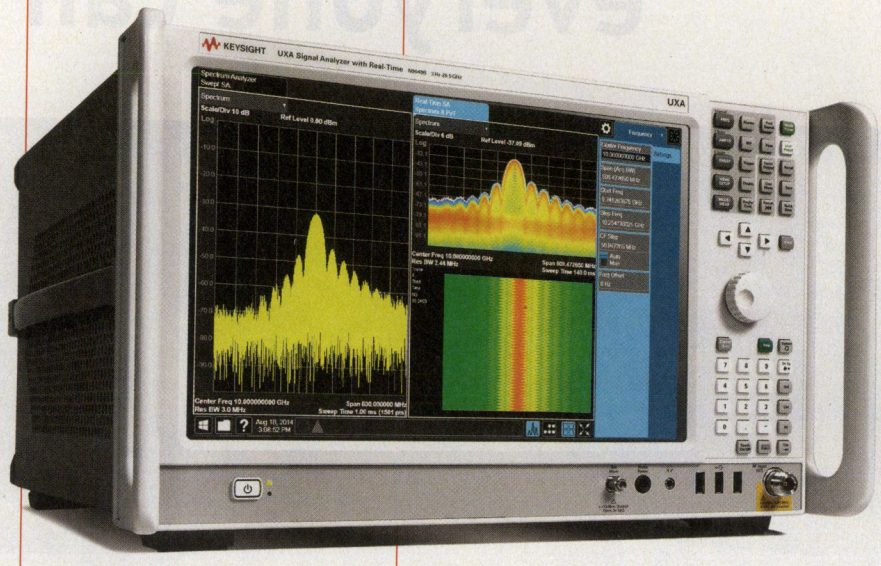

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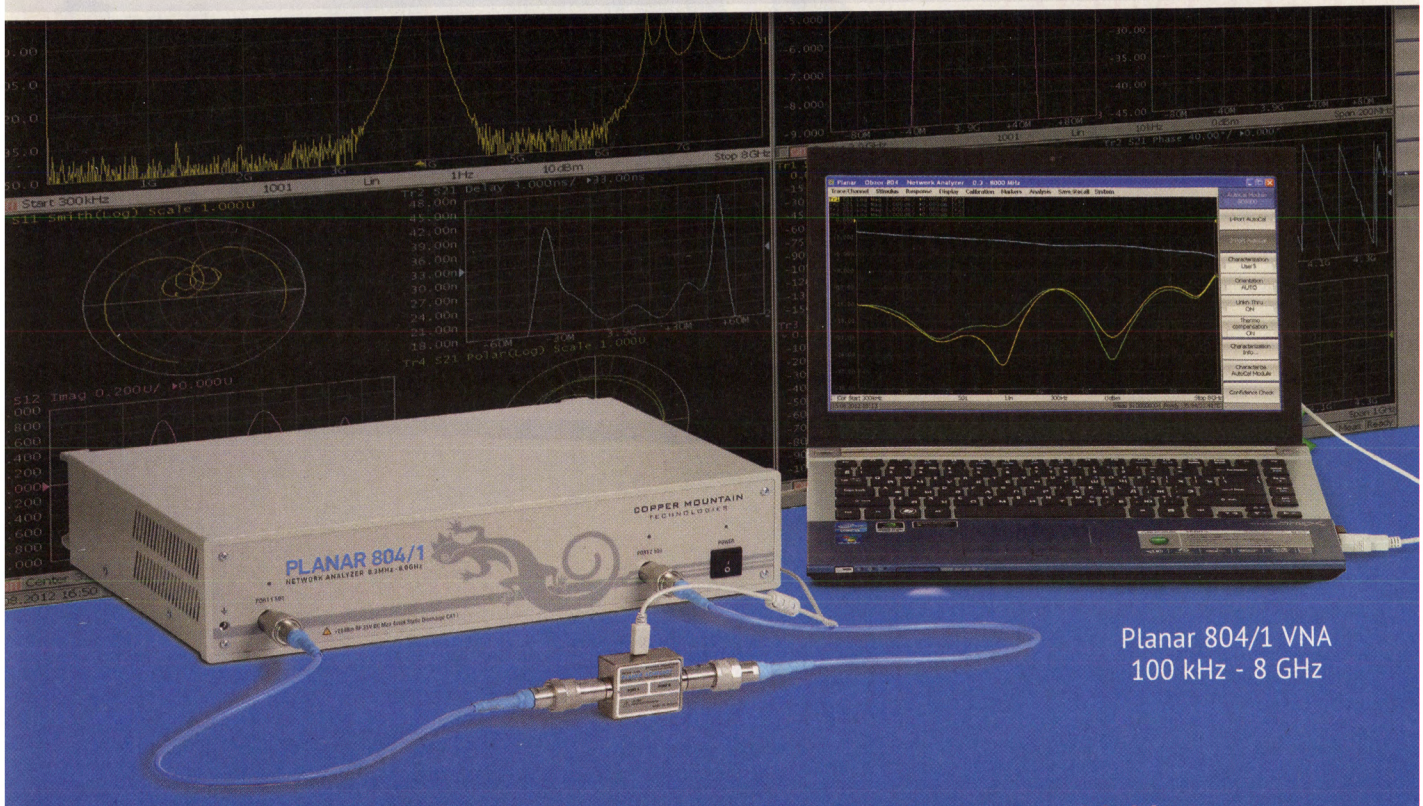

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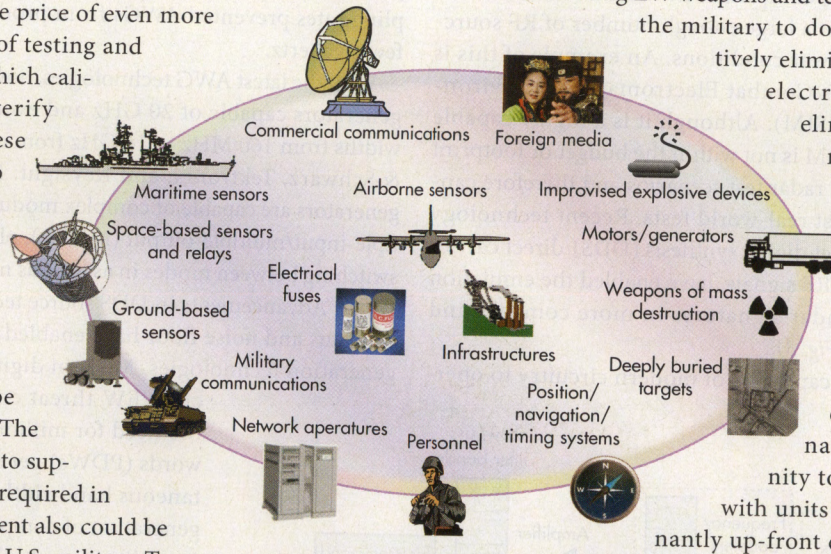
JEAN-JACQUES DELISLE | Technical Engineering Editor

Signal-Generation Advances Support Electronic Warfare's Evolution

As electronic threats grow and the electronic-warfare environment becomes more complex, new methods of generating and controlling EW signals are needed to maintain a modern war-force.

MODERN ELECTRONIC WARFARE (EW) makes use of the latest RF/microwave technologies to attack, defend, and enhance strategic operations. To enable these applications, highly complex and integrated computer/radio systems are needed to keep the military on the cutting edge. This advancement comes with the price of even more advanced methods of testing and signal generation, which calibrate, refine, and verify the operation of these highly capable radio weapons, armor, and tools.

If the military could deploy EW instead of kinetic weapons, the cost of engagements could be dramatically lowered. The necessary supply lines to support the ammunition required in a traditional engagement also could be cut. For example, the U.S. military Tomahawk (which The Economist claims cost $1.5 million per missile) and the Hellfire air-to-ground rocket (estimated at a cost of $115,000 per rocket) can now be used in large numbers during engagements. On September 23 of this year, the Washington Post reported that 47 Tomahawk missiles were fired overnight between two U.S. warships in the Persian Gulf and the Red Sea. Other examples can be found in the defense of U.S. assets—for example, the $70 billion price tag on the F-22 Raptor fighter aircraft program. Obviously, the use of such assets could be greatly reduced if it were possible to instead eliminate an opponent's ability to coordinate and communicate between forces while maintaining our forces' communications capability.



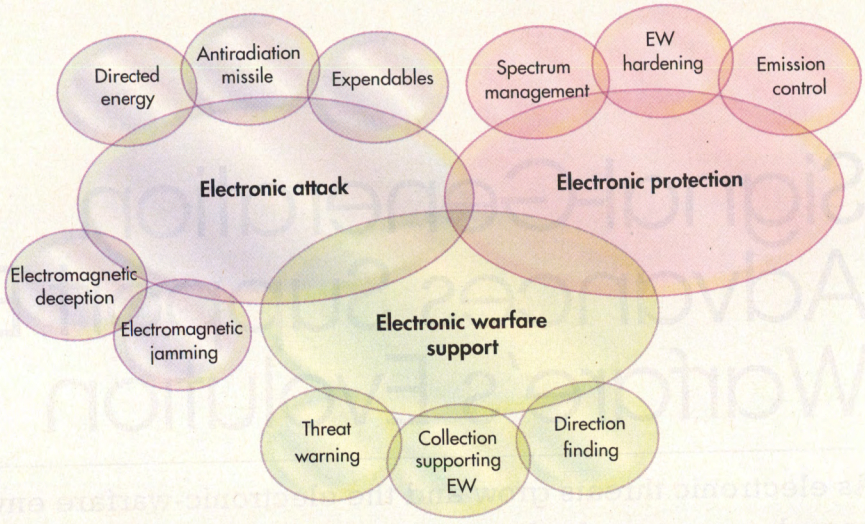
1. Any electronic or electromagnetic device/equipment on the battlefield, including military/industrial infrastructure, is subject to disruption and even elimination by EW attack. *(Courtesy of the Department of the Army)*

Substituting EW weapons and techniques could enable the military to do the following: effectively eliminate a target through electrical attack, reduce/eliminate the effectiveness of an opponent's attack, or strategically outmaneuver an opponent through superior communication or resource denial *(Fig. 1)*. Within each of these scenarios is an opportunity to implement RF EW with units that incur predominantly up-front expenses. For example, an active-electronically-steered-array (AESA)-based jammer may cost hundreds of millions of dollars initially to develop and millions more to deploy. Yet the residual costs may be minimal in comparison to current kinetic-weapons technology.

The U.S. Navy and other branches are investing heavily in technology to provide electronic attack, protection, and warfare support *(Fig. 2)*. To enable these technologies, software-defined radios (SDRs), cognitive networking, and radars based upon antenna arrays are being imple-

mented. These technologies have begun to utilize more cost-effective commercial-off-the-shelf (COTS) technology. They also take advantage of the latest high-performing materials, such as gallium nitride (GaN). These new technologies raise performance while adding a wide range of capabilities.

To take greatest advantage of these benefits, however, these devices require more advanced signal-generation technology. "The modern spectral threat environment contains thousands of emitters—radios, wireless devices, and tens to hundreds of radar threats—producing millions of radar pulses per second amidst background signals and noise," says Shawn McCann, EW Business Development Manager for Keysight Technologies. "Recreating this environment in the laboratory is a major challenge, often involving a number of different types of signal-generation tools."

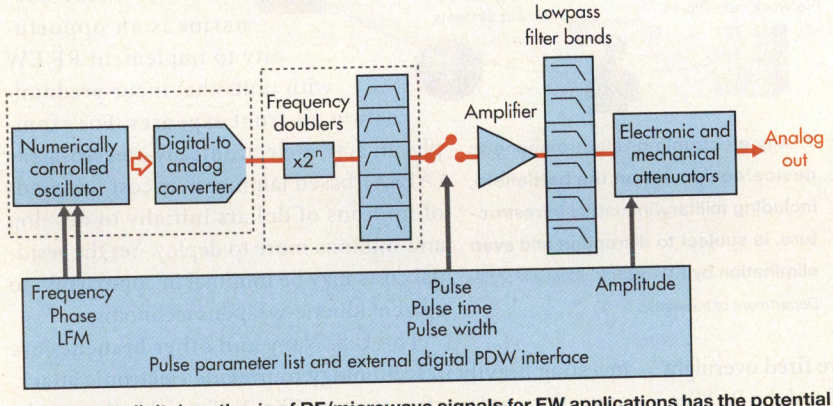


2. To modernize a military with EW capability, forces will need to be equipped with offensive, defensive, and strategic EW devices and techniques. *(Courtesy of the Department of the Army)*

Traditionally, large, complex rack-mount cabins would be necessary to present a large enough number of RF sources to emulate true-to-war conditions. An example of this is Northrop Grumman's Combat Electromagnetic Environment Simulator (CEESIM). Although it is a highly capable instrument, the CEESIM is not within the budget or footprint requirements of many radar test scenarios and therefore cannot be applied to most real-world tests. Recent technology advances, such as direct digital synthesis (DDS)/direct carrier generation (DCG) of RF signals, have enabled the emulation of highly complex radar signatures in more compact and cost-effective forms *(Fig. 3)*.

DDS leverages the capability of modern circuitry to operate at sampling rates that have reached RF frequencies. With an ideal arbitrary waveform generator (AWG), a signal up to half of the sampling rate can be produced with sufficient accuracy to enable directly generated, modulated RF. In the past, low spurious-free-dynamic-range (SFDR) and sampling rates prevented AWGs from producing greater than a few gigahertz.

With the latest AWG technologies, however, there are signal generators capable of 20 GHz and even 40 GHz and bandwidths from 160 MHz to 14 GHz from companies like Rohde & Schwarz, Tektronix, and Keysight. Many of these signal generators are capable of complex modulation including multiple-input/multiple-output (MIMO), phased array, and rapid switching between modes in as little as nanoseconds. McCann notes, "Advancements in DDS source technology with a lower spurious and noise floor have enabled the latest EW signal-generation technologies. Modern digital interfaces for efficient EW threat environments support the need for millions of pulse descriptor words (PDWs) per second. Wider instantaneous bandwidths are required for the generation of complex spectral scenarios. In addition, multiple synchronized channels are needed to simulate signals for systems using multiple or phased-array antenna systems."



3. The direct digital synthesis of RF/microwave signals for EW applications has the potential to generate highly complex EW signals in a cost and form factor that is significantly reduced when compared to traditional technologies. *(Courtesy of Keysight Technologies)*

Although material developments have increased circuitry performance, clever techniques are still being developed to significantly enhance the present technology for signal generation. An example of an advanced DDS technique is Tektronix's AWG70001 dual-DAC architecture, which

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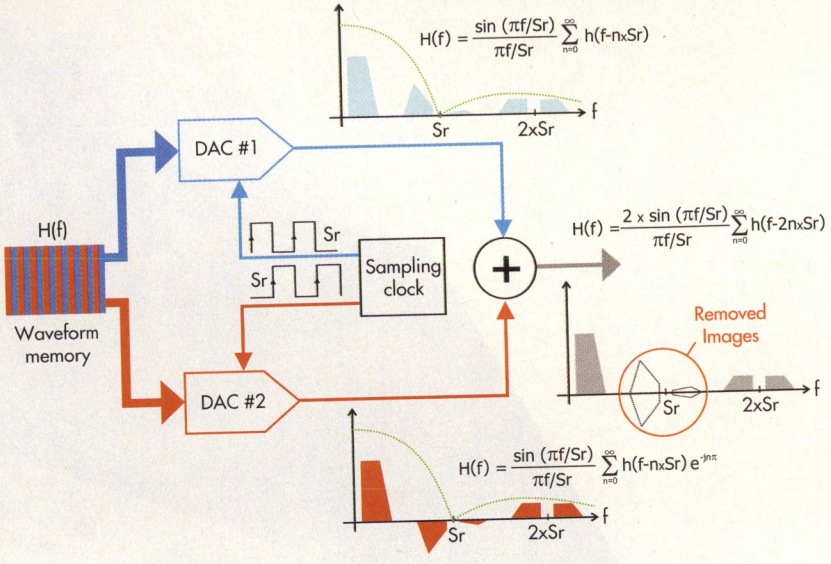
> "As many of the world's militaries move to electronic warfare as a means of reducing cost and casualties...the investment in high-performing/agile signal-generation technologies will continue to scale."

enables extended AWG range using two matched digital-to-analog converters (DACs) (*Fig. 4*). The first DAC has the odd samples delivered to it, while the second DAC has the even samples applied. A delay of half of the sampling period must be applied to the second DAC in order to cancel the images from each DAC. This approach prevents those images from appearing in the second Nyquist band. According to Tektronix, this approach enhances a DDS structure to run at two times the sampling rate instead of half the sampling rate of a single DAC architecture.

Yet there are limitations to DDS technology. The record length (RL) of a radar signal limits a time window in inverse proportion to the sampling rate. For higher sampling rates, the RL time window is therefore shortened. Yet the RL remains critical for simulating real-world and diverse radar signals and environmental factors. With a long RL, for example, the effects of physical vibration, acceleration, echo characteristics, frequency-hopping pattern, and staggered pulse sequences can all be included in simulation. The result is more effective radar calibration.



4. Using a slightly phase-shifted dual DAC architecture can artificially increase the sampling rate and replicate signals of much higher frequencies while reducing images. *(Courtesy of Tektronix)*

Other advances also have pushed the boundaries of EW signal generation. "With EW systems, the main technology advancements have been modern digital signal processing (DSP), higher-performance analog-to-digital converters/DACs, and software-defined system architectures," says McCann. "On the EW-simulator side, advancements have been enabled by core DAC performance, nano-field-effect-transistor switching, 40-GHz agile attenuators, real-time pulse generation, Ethernet interfaces, simulator angle-of-attack calibration/verification methods, and threat reactivity based on behavioral models."

Cognitive-networking technologies also may be applied to the modern battle space. According to Exelis, for instance, its Disruptor SRx, which uses cognitive EW technology, can respond in real time to an array of previously unknown waveforms and adapt to new operating modes. This technology is designed for use cases ranging from unmanned aerial vehicles (UAVs) to combat aircraft. It uses a systems-of-systems (SoS) approach.

With digital RF memory, DSP, and a digital receiver incorporated into the same module, the Disruptor SRx may be capable of using traditional threat libraries in addition to classifying unknown signals and generating jamming using advanced machine learning in real time. Such methods can remove the human-manipulation aspect from the characterization of EW signals. With the benefit of an intelligence node, they also can enable EW systems to react without hesitation in virtually any battlespace condition, regardless of disrupted communications.

As many of the world's militaries move to electronic warfare as a means of reducing cost and casualties, and increasing response time, the investment in high-performing/agile signal-generation technologies will continue to scale. The testing of modern EW devices is scaling with this increasing complexity.

This trend is further driving the demand for increased frequency, modulation capability, bandwidth, and signal fidelity from test-equipment waveform generators. Inevitably, hardware-in-the-loop and automated testing techniques using highly integrated and computerized software-defined systems will be crucial to the proper design and "training" of the latest radars, jammers, and support networks. MW

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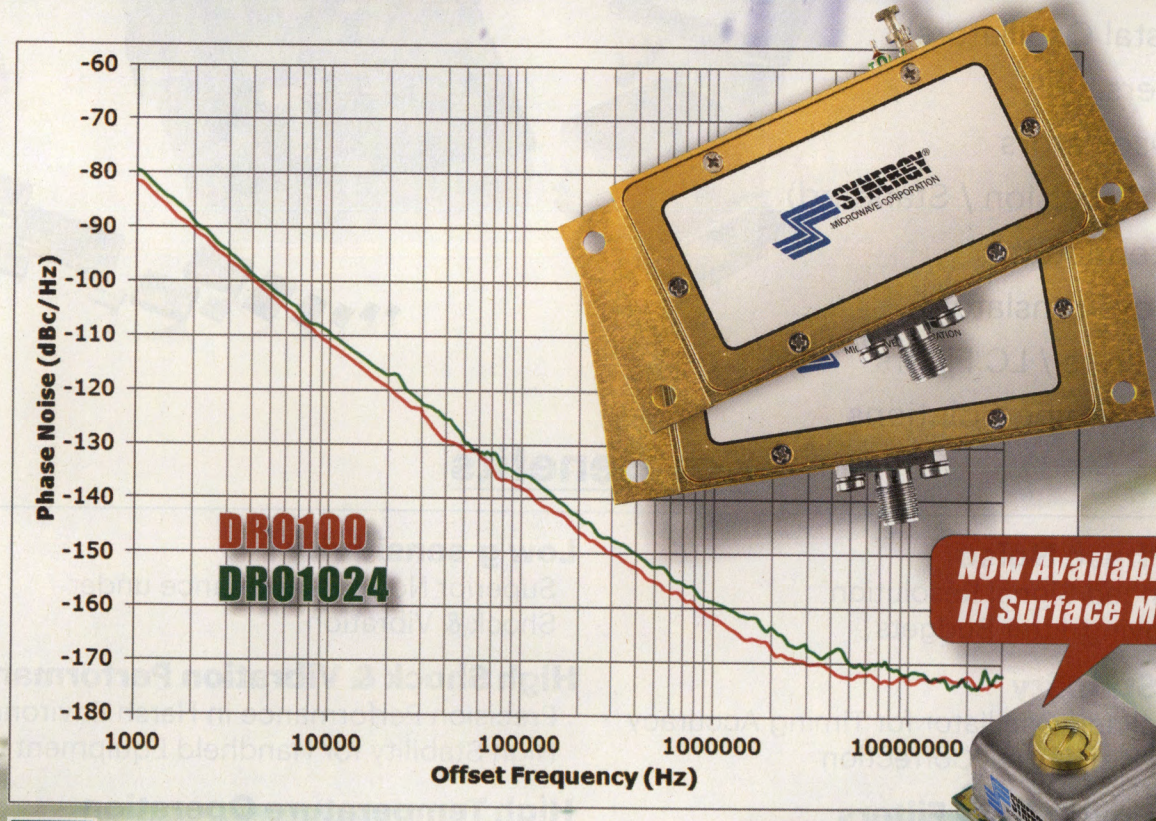

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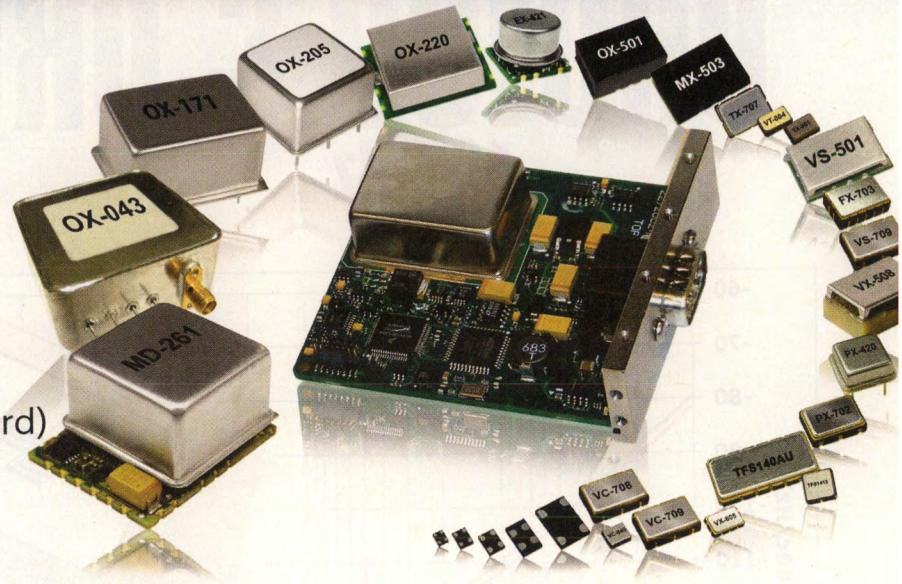

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A Short History of HARMONIC LOAD PULL

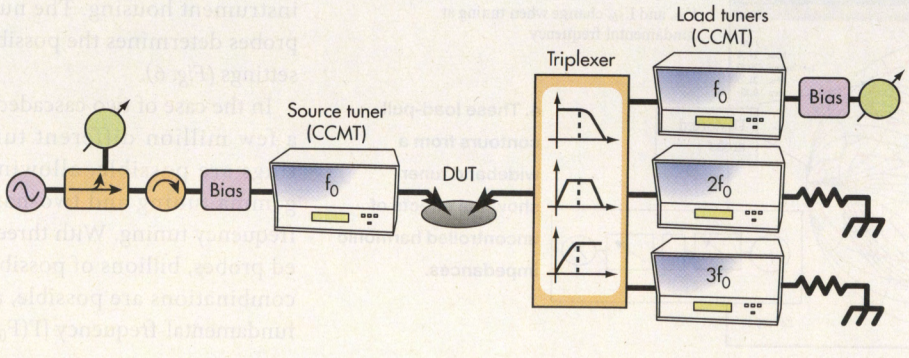
Harmonic load-pull tuners are important systems for characterizing power transistors and amplifiers and finding the impedances needed for gaining optimum performance levels.

Load-pull tuning has long been a vital tool in analyzing and characterizing active devices, such as RF/microwave power transistors, to changes in the source and load impedances presented to the device. Typically, load-pull tuners change the impedances to the active device while additional measurement equipment—like a vector network analyzer (VNA)—is used to measure device characteristics, such as amplitude, phase, and scattering (S) parameters.

By tuning the impedances during measurements, the optimum source and load impedances to the device under test (DUT) can be found for the creation of impedance-matching circuitry. This enables transitions from the device impedances to the characteristic impedance (typically 50 Ω) of the circuit or system in which the active device will be used.

Over the years, load-pull technology has evolved from its humble beginnings to more elaborate and sophisticated tools and techniques. Many load-pull tuners developed with broadband capabilities and intended for creating controlled impedances at fundamental frequencies, or F_0, have also created uncontrolled impedances at harmonic frequencies of the fundamental, such as $2F_0$, $3F_0$, and so on. Most engineers working with load-pull testing have worked with such an uncontrolled harmonic test environment for at least a 15-year period.

Efforts to control this test environment led to the development of harmonic load-pull solutions, using a fundamental-frequency test tone and one or more harmonic frequencies of that fundamental frequency. The use of harmonic load-pull testing has often helped to increase the power-added efficiency (PAE) of an amplifier or active device, or else reduce the error vector magnitude (EVM) of a power



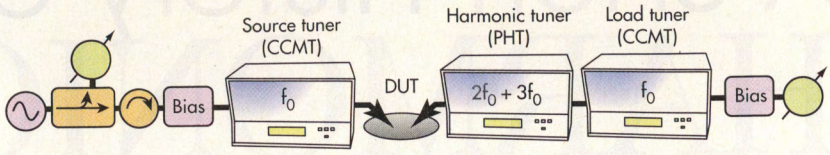
1. Early harmonic load-pull tuners typically employed multiplexers to control harmonic levels.

> "The MPT concept apparently works for all frequencies and impedances within a given tuning range and within the common frequency bandwidths of all probes."

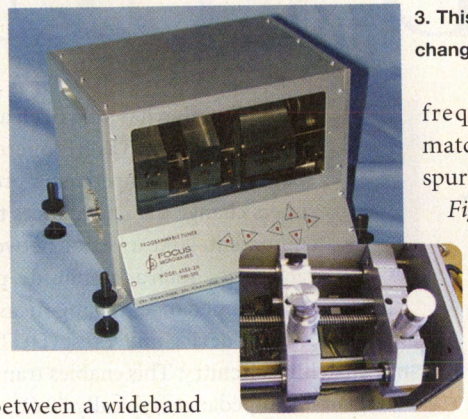
amplifier being operated in backed-off bias conditions.

Early attempts to develop harmonic load-pull tuners typically inserted narrowband frequency discriminators, such as diplexers or triplexers, between a DUT and a wideband impedance tuner *(Fig. 1)*. While such an approach can provide some control over the harmonic frequencies, it is typically limited in its power range and bandwidth and susceptible to spurious signal products.[1]

By 1998, a simple solution had been proposed for load-pull harmonic tuning,[2] which was "productized" in the form of the Programmable Harmonic Rejection Tuner (PHT) from Focus Microwaves in 2000. *Figure 2* shows a load-pull setup using a harmonic-rejection tuner on the output of the DUT.



2. The use of harmonic rejection tuners enabled load-pull testing with some degree of harmonic control.



3. This is a PHT and its internal mechanism for changing resonators.

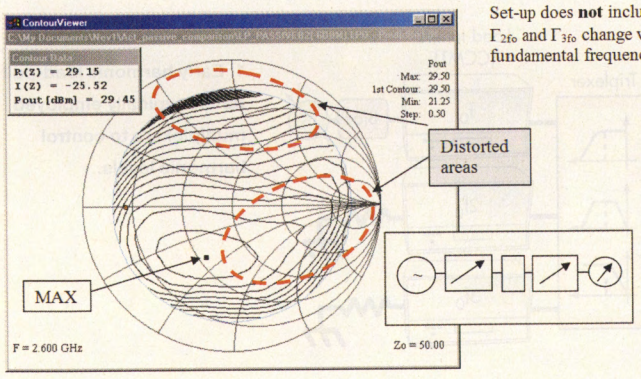
These harmonic tuners are inserted between a wideband tuner and the DUT. They use a set of remote-controlled open-stub quarter-wavelength resonators at the first- and second-harmonic frequencies ($2F_0$ and $3F_0$). They slide on top of the center conductor of the airline used for the wideband load-pull tuner and can be tuned back and forth with little ill effect on the fundamental-frequency impedance, $\Gamma(F_0)$.

The approach used in the PHT system can provide full 360-deg. phase control at high values of Γ on the order of 0.98, critical for optimizing the PAE of a power amplifier. The PHT systems have two important characteristics: their resonators reflect all harmonic power back to the DUT so that the wideband load-pull tuner acts as a pure fundamental-frequency tuner, and they are 50-Ω-matched at low frequencies to minimize spurious oscillations.

Figure 3 shows a PHT and its internal mechanism for changing resonators. *Figure 4* depicts load-pull contours from a wideband tuner (the deformation due to uncontrolled harmonic impedance can be seen). *Figure 5* shows load-pull contours measured after a PHT has been inserted to maintain constant harmonic impedances.

The PHT systems enabled analysis of high-Γ load-pull contours for many PHT users, providing data under controlled-impedance test environments.[2,3] Still, these systems suffered frequency-band limitations and the amplitude of Γ could not be controlled, leading many customers to desire more wideband tuning and full control of Γ at harmonic frequencies.

For versatility, the Multi-Purpose Tuner (MPT) was introduced in 2004 (United States patent No. 7,135,941), with the first units appearing on the commercial market in 2006. The MPT is essentially a cascade of three wideband, single-probe tuners, all covering the same bandwidth and contained within a single instrument housing. The number of probes determines the possible tuner settings *(Fig. 6)*.



Set-up does **not** include **harmonic** tuners: Γ_{2fo} and Γ_{3fo} change when tuning at fundamental frequency

4. These load-pull contours from a wideband tuner show the effects of uncontrolled harmonic impedances.

In the case of two cascaded probes, a few million different tuner settings are possible, allowing high-gamma tuning and two-harmonic-frequency tuning. With three cascaded probes, billions of possible probe combinations are possible, allowing fundamental-frequency [$\Gamma(F_0)$] and as



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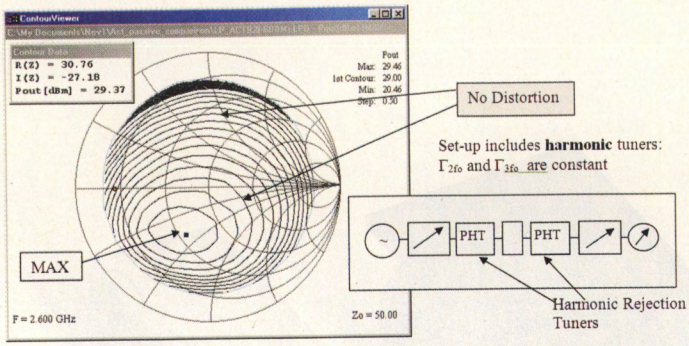
many as three harmonic reflection factors [$\Gamma(2F_0)$, $\Gamma(3F_0)$] to be synthesized simultaneously and independently.

The MPT concept apparently works for all frequencies and impedances within a given tuning range and within the common frequency bandwidths of all probes. A two-probe impedance tuner can cover the entire Smith chart for two frequencies, and can also tune three frequencies independently, but it cannot do so over the entire Smith chart. There are areas of the Smith chart, inside the $|\Gamma|$ range of the probes, which harmonic tuning simply cannot reach.



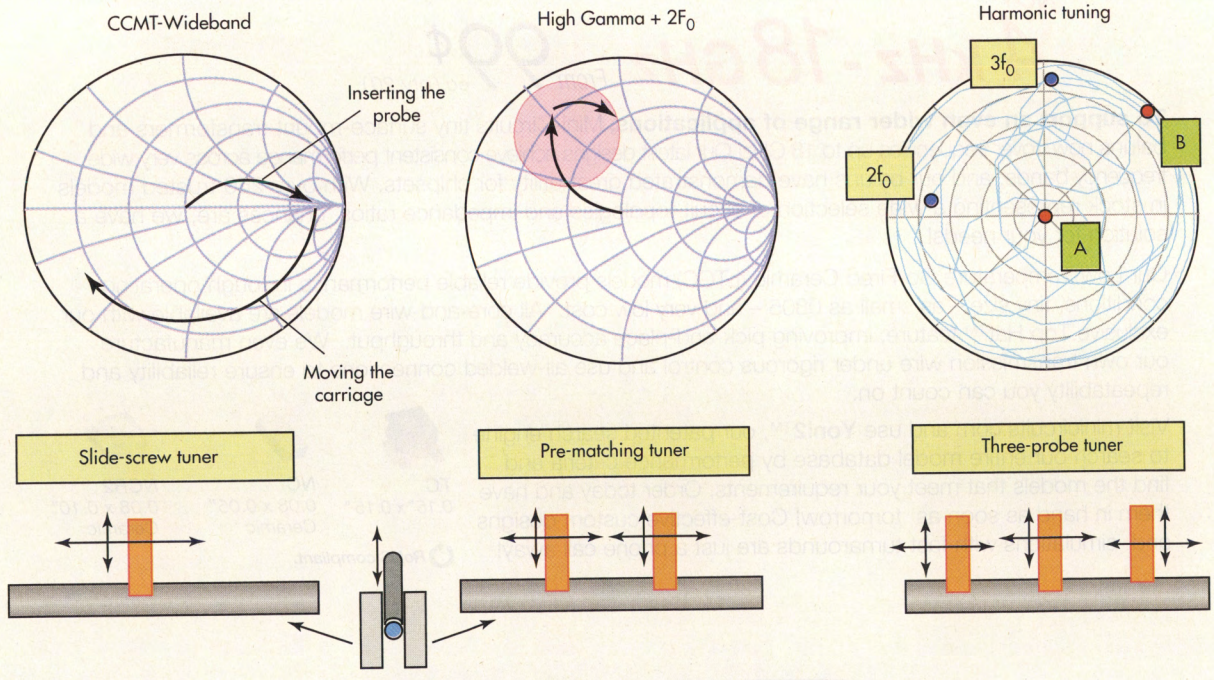
5. These are load-pull contours measured after a PHT system has been inserted to maintain constant harmonic impedances.

Given the limitations of the two-probe load-pull systems, integrating three probes within a single load-pull system housing (using the same airline) made sense as part of the evolution of load-pull technology. Such as approach, implemented in three-probe MPT systems, eliminates adapter loss and residual reflections associated with assemblies of three cascaded wideband tuners.

A system with the three integrated probes must be calibrated within a limited period of time, and developing such a system requires the development of efficient tuning algorithms using the calibration data. Unfortunately, calibrating each of the billions of possible probe positions could possibly take years. Part of patent No. 7,135,941 is a calibration method. It involves de-embedding of the initialized impedance tuner—a concept not immediately obvious, and one which enables calibrations of MPT systems in just minutes.

The second challenge in designing an effective three-probe harmonic load-pull system (efficient tuning and high harmonic isolation) required several years of development and many engineering iterations, as is often the case for original designs (patent Nos. 8,629,742, 13/915,160, and 12,929,643, pending). Admittedly, the first harmonic load-pull tuning operation in 2004 with a three-probe MPT system took more than 30 min, a task which today requires barely more than two seconds.

Achieving high tuning accuracy and high harmonic tuning accuracy can be extremely difficult for a system housing a cascade of three independent, highly reflective probes and the same airline. But after almost 10 years of work on this three-probe concept at Focus, the MPT system approach provides



6. A three-probe load-pull tuner system allows an almost infinite number of different probe position combinations.

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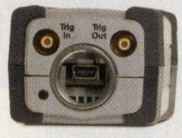

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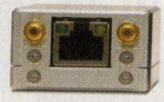

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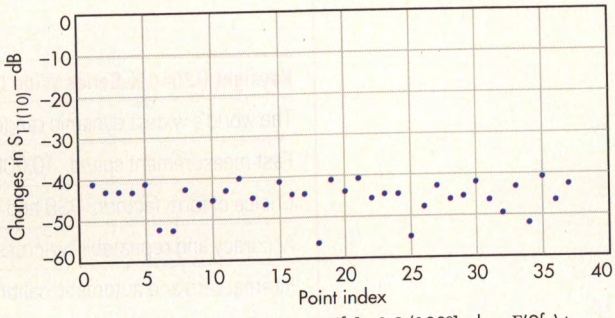

7. The MPT system delivers high harmonic tuning isolation with tuning accuracy of 50 dB or better.

Pattern calibration: Change in $\Gamma(f_0[= 0.9/180°]$ when $\Gamma(2f_0)$ tuners on a load-pull wedge 0-0.9/120-190°

excellent tuning accuracy of typically 50 dB or more and similar harmonic tuning isolation *(Fig. 7)*.

Gaining these excellent performance levels required ongoing improvements in software algorithms and mechanical precision. While a single-probe tuner may have possibly 10 problems or issues for a developer, a three-probe MPT system with its greater complexity can present as many as 1000 issues to overcome.

The validity of a four-probe harmonic load-pull tuning approach was explored by designing a tuner named "Quattro" for operation from 1.8 to 10.0 GHz. It is capable of independently tuning across frequencies from the fundamental, F_0, to the fourth harmonic at $4F_0$. In measurements made with this four-probe tuner, all points of a load-pull pattern at a particular fundamental frequency, F_0—such as the 2 GHz in *Fig. 8* (the black dots)—can be tuned while $\Gamma(2F_0)$ to $\Gamma(4F_0)$ remain fixed (the colored dots in *Fig. 8*).

ACTIVE TUNING

The use of active tuning offers several potential benefits in the evolution of load-pull tuners (e.g., overcoming the fixture insertion loss limiting passive load-pull tuners). One definition of active device tuning refers to injecting into the output of a transistor a signal which is coherent or synchronized with the input signal and can be controlled in terms of amplitude and phase, to evaluate a virtual or active load. For the transistor, this would be as if the signal was reflected by a passive tuner. In addition to the loss compensation leading to the fact that $|\Gamma| = 1$ can be reached, the active tuning approach can be implemented with standard commercial components, bypassing the mechanical precision required for fabricating passive electromechanical impedance tuners.

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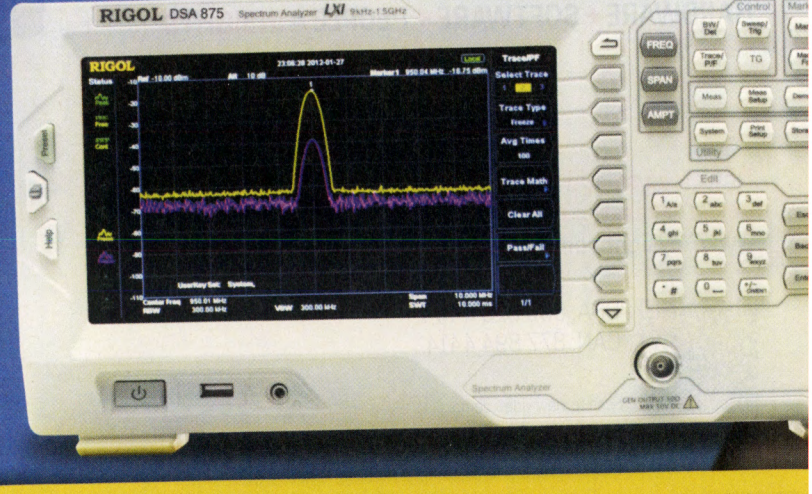

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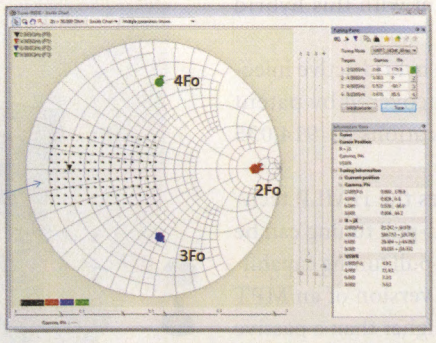
Active load-pull tuners can be designed as closed-loop or open-loop systems. In closed-loop active load-pull systems, the outgoing signal is sampled, amplified, phase modulated, and re-injected into the transistor output. Technical details on such tuner approaches have been published since the mid-1980s, with a commercial product, the Active Load Pull System (ALPS), introduced by Focus in 1998.[7] But because of spurious signal issues and other side effects, these active load-pull systems did not make any type of penetration into the RF/microwave market.



8. These plots are from a four-probe load-pull system, with the black dots representing fundamental impedances and the colored dots showing harmonic impedances.

Open-loop active load-pull systems, where a second synchronized signal is injected into the transistor output, avoid many of the problems associated with closed-loop load-pull systems, such as spurious oscillations. However, a shortcoming of the open-loop approach is the need for high RF/microwave power of the output-injected signal source or power amplifier, to overcome the impedance mismatch between the low impedance of many power transistors (about 1 Ω in the setup and the internal impedance of the injecting signal source (50 Ω).

At very high-Γ values, the required injection power can reach 20 times the output power of the transistor under test. This simply may not be possible or financially affordable. This high power requirement can be relieved by using an impedance transformer (pre-matching tuner) between the transistor and the load—i.e., configuring a "hybrid" or "active-passive" tuning system. This reduces the need to inject power roughly by a factor of 10. In a hybrid system, the injection power requirement is reduced to less than two times the transistor output power.

As far as the tuning speed limitations of mechanical load-pull tuners, a hybrid load-pull system can be configured with the mechanical tuner placed permanently in a position optimized for overall reduced injection power. All remaining tuning is then performed electronically and quite quickly, depending essentially on the response time of the receiver (load-pull patterns including 50 points have been

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measured in roughly 1.5 s or 35 ms/point with this hybrid load-pull tuner approach).

This hybrid solution is promoted as the hybrid active injection load pull (HAILP) technique. It is covered by patent No. 8,497,689; as well as pending applications 61/914,035 and 62/042,550.

If an MPT system *(Fig. 9)* incorporates the HAILP technique, independent passive harmonic load-pull tuning can be overlapped with the active harmonic load-pull tuning capability of the MPT system. In such a hybrid version of an MPT system *(Fig. 10)*, the optimum Γ value is higher than a passive system, (close to $|\Gamma| = 1$), and the maximum output power of the hybrid system is almost 5 dB higher than the passive version.

The hybrid system offers an additional benefit for verifying the overall system accuracy: At the same values of passive (pre-calibrated) impedance and active impedance, the system must generate the same output power and gain. This can easily be done by switching between purely passive and purely active tuning to the same load impedance. Having higher Γ values lead to higher PAE performance in a DUT and a higher PAE swing.

Harmonic load-pull tuning has progressed a great deal since the introduction of automatic impedance tuners in 1985. Innovations have enabled progress, but some good ideas may not always be practical, depending upon available technology. For example, the MPT tuning algorithms now used would not have been possible in 1998 when the PHT was introduced: the necessary computing power for those algorithms simply was not affordable for the majority of working engineers in 1998.

Not all ideas can be turned into practical solutions, but the innovations from those working in the industry on load-pull systems may hopefully continue to surprise and serve those who can benefit from those systems.[14,15] mw



9. The MPT-5080 load-pull tuner is an example of a system designed using patented active-passive tuning technology.

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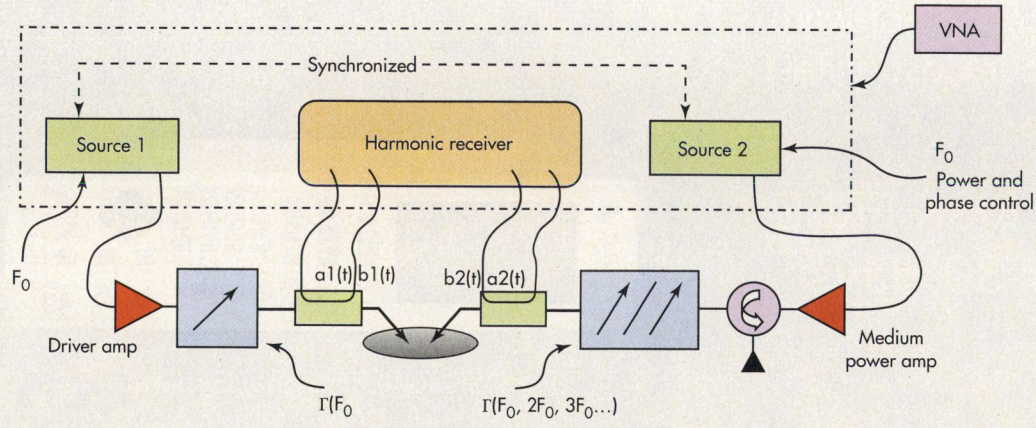

10. The patented HAILP active-passive load-pull tuning techniques provide flexibility with high accuracy across microwave and millimeter-wave frequencies.

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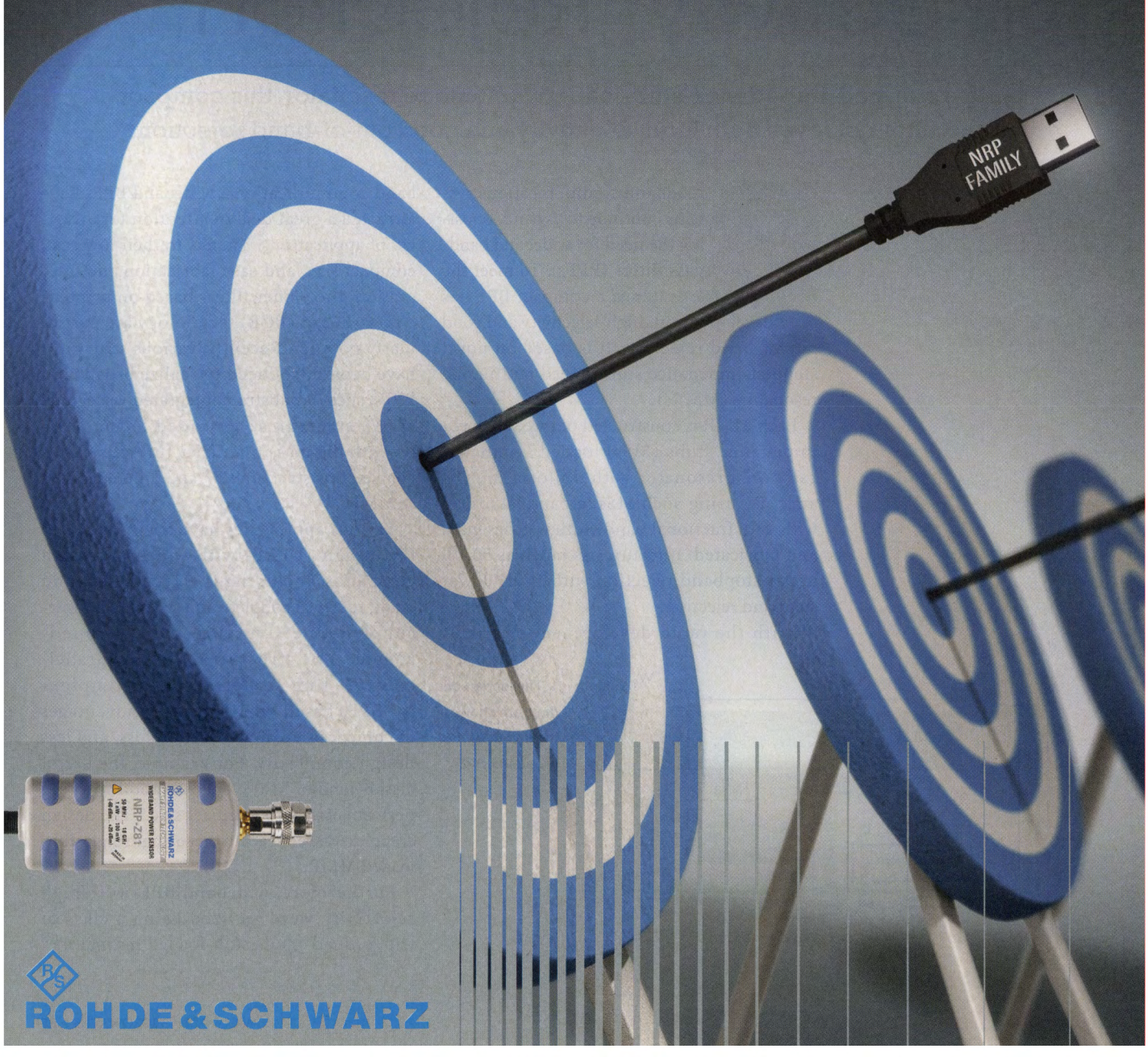

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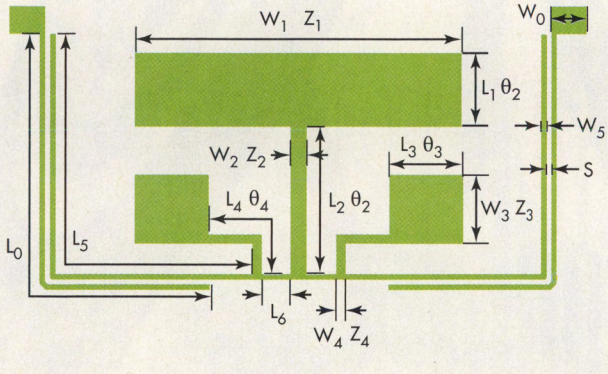
Bandpass Filter Combines High Selectivity, Rejection

By using a stepped-impedance stub-loaded multimode resonator, this compact wideband filter achieves both high selectivity and good out-of-band rejection.

Broadband communications systems continue to expand, boosting the need for wideband bandpass filters (BPFs). To meet the needs of those systems, a compact BPF has been developed with high selectivity and good out-of-band rejection. It is based on novel stepped-impedance stub-loaded multimode resonators (MMRs).

Each MMR is constructed of three stepped-impedance stubs (SISs) loaded symmetrically on a resonator with uniform impedance. By using such SISs, a bandpass filter with 48% fractional bandwidth was designed and fabricated. It features as much as 50-dB upper stopband rejection with 35-dB lower stopband rejection.

With the rapid development of broadband communications, wideband BPFs have attracted a great deal of attention for a variety of applications. Thanks to their low cost, compact size, and easy fabrication and integration, the planer filters based on printed-circuit-board (PCB) technology are particularly popular. Recently, various structures have been introduced to realize broadband BPFs, such as substrate integrated waveguide (SIW) structures,[1] cascaded lowpass filter with highpass filter structures,[2] parallel-coupled line structures,[3-5] and different kinds of MMRs.[6-12]



1. This schematic diagram shows the layout of the proposed BPF.

Various approaches have been applied to the design of BPFs with wide-bandwidth passbands, including the use of a ground plane aperture coupled-line section,[3] a three-coupled-line structure,[4] and stepped-coupled-line structure,[5] but the parallel-coupled-line structures add complexity. BPFs based on MMRs offer advantages of compact size, low insertion loss, and less design complexity. For example, the use of a stub-loaded MMR enables good filtering performance with compact size.[6] The design complex was reduced further using a single stub-loaded MMR.[7]

Furthermore, wideband BPFs with high selectivity were realized using a class of triple-/quad-mode stub-loaded resonators.[8]

A compact, high-rejection wideband BPF with improved upper stopband was realized with a quad-mode stub-loaded resonator.[9] Also, a stub-loaded MMR was used for a BPF with wideband and high rejection,[10] while a wideband BPF with high rejection was realized with a folded stepped-impedance resonator.[11]

Reference 12 details a stepped-impedance resonator loaded with a uniform-impedance microstrip stub to design a sharp-rejection broadband BPF. Unfortunately, most of the stub-loaded MMRs are constructed by loading constant-impedance stubs, which cannot achieve high selectivity and good out-of-band rejection simultaneously.[6-12] Low-impedance stubs support BPFs with good out-of-band rejection but poor selectivity.[10-12] In contrast, high-impedance stubs enable BPFs with high selectivity but poor out-of-band rejection.[7-9]

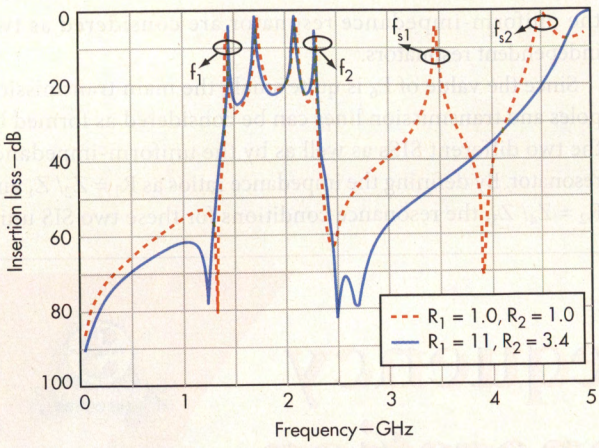
To achieve a wideband BPF with high selectivity and good out-of-band rejection, a design was developed with a stepped-impedance stub-loaded resonator. The filter's MMR is constructed by loading three SISs symmetrically on a uniform-impedance resonator. The resonator itself can provide four transmission poles under weakly coupled conditions and six transmission poles under strongly coupled conditions.

Two transmission zeros are obtained beside the upper and lower edges of the passband, helping to significantly improve its selectivity. Rather than uniform-impedance stubs, SISs are used for higher selectivity and better rejection performance over wide stopbands, compared with the conventional wideband BPFs based on stub-loaded MMRs in refs. 6-12.

Figure 1 presents the layout of the proposed BPF. The MMR is constructed by loading one SIS on the center plane and two SISs symmetrically on a uniform-impedance resonator with small dimension L_6. According to the analysis in ref. 6, the open-ended stub loaded on the center plane of a uniform-impedance resonator works as a K-inverter between the two resonators, where the open stub and one-half of the uniform-impedance resonator are considered as two independent resonators.

Since the value of L_6 is quite small, the main transmission poles and transmission lines can be considered as formed by the two different SISs as well as by the uniform-impedance resonator. By defining the impedance ratios as $R_1 = Z_2/Z_1$ and $R_2 = Z_4/Z_3$, the resonance conditions for these two SIS units





2. Passband insertion loss has been simulated under weakly coupled conditions with varied impedance ratios R_1 and R_2.

can be expressed as[13]:

$$1/R_1 = \tan\theta_1(\tan\theta_2) \quad (1)$$

$$1/R_2 = \tan\theta_3(\tan\theta_4) \quad (2)$$

Figure 2 shows simulated insertion loss under weakly coupled conditions with variations in impedance ratios R_1 and R_2. The fundamental resonance frequencies of the two SISs are defined as f_i (i = 1, 2), while the first spurious resonance frequencies are defined as f_{si} (i = 1, 2). Four transmission poles were obtained with weak coupling while as many as six transmission poles can be obtained with strong coupling due to parasitic coupling effects. Only the transmission poles and zeros near the passband edges need be analyzed to reconfigure the cutoff frequencies for this BPF design.

The filter's transmission pole (f_1) and the transmission zero located beside the lower edge of the passband are obtained by means of the SIS in the center plane, while the transmission pole (f_2) and transmission zero located near the upper edge of the passband are results of the sym-

SCREENING THROUGH BANDPASS-FILTER SPECS

BANDPASS FILTERS MAY be among one of the more important components in a communications receiver, even though they are easily overlooked and quite often taken for granted. While newer structures such as multimode resonators (MMRs) are being recruited to create bandpass filters with increasingly wide fractional bandwidths, there are still strong needs for RF/microwave bandpass filters with conventional fractional bandwidths and frequency ranges, and for a clear understanding of the important performance parameters to evaluate when designing and/or choosing a bandpass filter for a particular application.

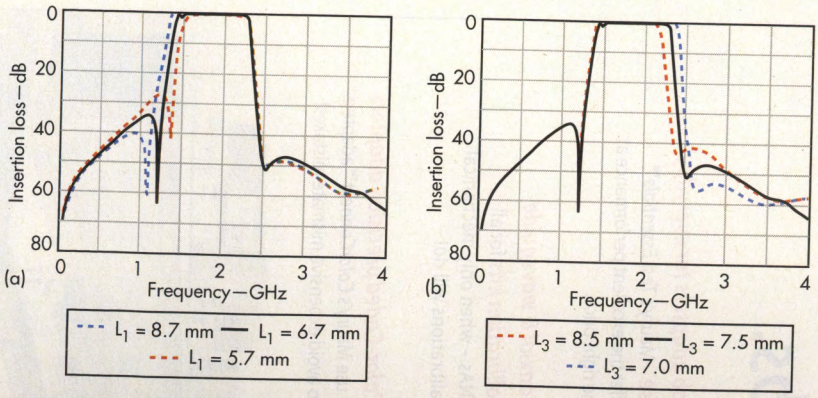
Any filter can be thought of as a form of tuner. Just as an old-fashioned AM or FM radio could be controlled via a rotary dial when "tuning" or searching for a particular radio station, an RF/microwave filter is used as a fixed tuner. In both cases, many signals are present simultaneously in the environment, but only one signal or, in the case of a filter, perhaps a handful or specific frequency band of signals is of interest. An effective bandpass filter, as its name suggests, passes a designated frequency band of signals with as little loss as possible, while stopping all of the other signals outside of that frequency band.

A kind of inverse to the bandpass filter, the bandstop filter, does the reverse, in being designed to stop or attenuate as much as possible a specific frequency band of signals and pass the signals around that stopband with as little loss or attenuation as possible. As is apparent by the AM/FM radio example, and the growing number of wireless communications and other radio signals in the environment, such filters are vital to effectively maintaining modern communications systems.

BASIC PARAMETERS

Bandpass filters are typically characterized by a center frequency within their passband, usually designated at f_0. Depending on the needs of an application, the passband may be designed for narrowband or wideband use and the passband is some spectral distance on each side of the center frequency, such as $f_0 \pm 10$ MHz. Outside of the passband, signals are rejected or attenuated by some amount, such as 40 dB or −40 dBc at $f_0 \pm 10$ MHz. In addition to bandpass and bandstop filters, RF/microwave systems rely on lowpass and highpass filters. Lowpass filters are designed to pass with minimal loss all signals below a particular cutoff frequency and apply attenuation or rejection to all signals above the cutoff frequency. Highpass filters are designed to pass with minimal loss all signals above a specific cutoff frequency and attenuate or reject all signals below the cutoff frequency.

Bandpass and other filter types are available with different response types, such as Butterworth, Chebyshev, and Bessel filters. Each filter type exhibits a somewhat different amplitude response with frequency and each provides somewhat different transitions between passbands and stopbands. For example, the transitions from a bandpass filter's passband to its upper and lower stopbands can be rapid or they can be gradual, as typically describes by the different filter response types. Of course, there are always tradeoffs between one part of a bandpass filter's response and another, and finding the best filter for a particular application is usually a matter of making such tradeoffs work within that application. For example, Butterworth bandpass filters capable of extremely flat passband amplitude response with frequency, but typically lack sharp transitions from the passband to the stopbands. A Chebyshev bandpass filter can deliver those sharp transitions from the passband to the stopbands, but will typically exhibit much great passband amplitude ripple.



3. These plots of simulated insertion loss show the effects of adjusting parameters (a) L_1 and (b) L_3.

metrical SISs. *Figure 3* shows simulated passband insertion loss with changes in filter dimensions, the better to understand the influence of the two types of SISs on the BPF performance. The solid lines represent reference filters.

As *Fig. 3(a)* indicates, the lower cutoff frequency can be controlled by changing the dimension of the SIS in the center plane. As L_2 is varied from 8.7 to 5.7 mm, the lower 3-dB cutoff frequency increases from 1.32 to 1.50 GHz while the bandwidth of the filter tends to decrease. As *Fig. 3(b)* shows, the upper 3-dB cutoff frequency can be adjusted

The response of a bandpass filter will also be affected by how well it matches to the impedance of an application, which is usually 50 Ω. Filter data sheets list voltage standing wave ratio (VSWR) parameters to indicate how well a filter will match the impedance of a given system (such as 50 Ω in RF/microwave systems and 75 Ω in video systems), with lower VSWR numbers indicating better matching.

In general, the requirements of an application will define the necessary specifications for a bandpass filter, even in terms of package type. Stationary applications, such as cellular base stations, may require the higher power-handling capability of a larger bandpass filter, as achieved by filters with coaxial connectors. Portable applications, such as cellular telephone handsets, which do not have the large power-handling requirements of a base station, will benefit from bandpass filters in smaller packages, including chip-on-board packages and surface-mount-technology (SMT) packages. Bandpass filters are usable in both receive and transmit applications, and are available as both fixed and tunable filters, in which the passband can be shifted by some amount to meet the needs of an application. In the end, the application will set the performance levels that must be met by a particular bandpass filter.

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TTA1800-30-HG	1–18	45	3.0	3.0/4.0	10	2.5:1	EAR99
TTA1840-35	18–40	35	3.5	3.5	5	2.5:1	3A001b.4.c
TTA1840-35-HG	18–40	45	3.5	3.5	8	2.5:1	3A001b.4.c
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by altering the dimensions of the symmetrical SISs. As L_3 is varied from 8.5 to 7.0 mm, the upper 3-dB cutoff frequency increases from 2.16 to 2.38 GHz and the filter bandwidth tends to increase. By adjusting L_2 and L_3, the BPF's upper and lower cutoff frequencies can be tuned independently.

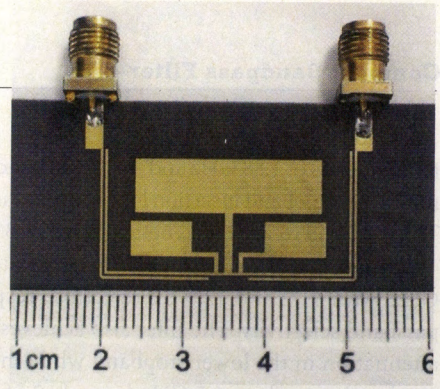
As for the upper stopband performance, the spurious frequencies of the SISs shown in *Fig. 2* can be expressed as Eq. 3, according to Eq. 11 in ref. 13:

$$f_{si}/f_i = \pi/[2\tan^{-1}(1/R_i)] \quad (3)$$

When $R_i > 1$, as the value of R_i increases, the spurious frequency tends to increase as well. Based on simulated results under weak coupling conditions, the impact of R_i and R_2 on the upper stopband of the proposed BPF can be seen in *Fig. 2*. When $R_i = 1$, the out-of-band performance in the upper stopband is quite poor due to the influence of spurious frequencies f_{s1} and f_{s2}. By properly adjusting the impedance ratios and dimensions of the SISs, a wideband BPF with high selectivity and good rejection performance can be designed.

Based on the analysis, a wideband BPF was designed and optimized. The dimensions of the BPF shown in *Fig. 1* were determined as $W_0 = 2.2$ mm; $L_0 = 29$ mm; $W_1 = 23$ mm; $L_1 = 6.7$ mm; $W_2 = 1$ mm; $L_2 = 7.35$ mm; $W_3 = 4.35$ mm; $L_3 = 7.5$ mm; $W_4 = 0.55$ mm; $L_4 = 5.2$ mm; $W_5 = 0.3$ mm; $L_5 = 28.42$ mm; $L_6 = 0.73$ mm; and $S = 0.3$ mm. The filter was fabricated on a commercial circuit-board substrate with dielectric constant of 2.55 in the z axis (thickness), thickness of 31 mils, and loss tangent of 0.0022. The effective area of the final BPF was $0.26\lambda_g \times 0.13\lambda_g$, where λ_g is the guided wavelength at f_0 the center frequency of the proposed BPF.

Figure 4 is a photograph of the fabricated BPF. It was characterized with the aid of a commercial vector network analyzer (VNA) from Keysight Technologies. The measured performance of the BPF is compared with simulated performance in *Fig. 5*, with close agreement. The measured 3-dB passband ranges from 1.43 to 2.33 GHz, with a cen-



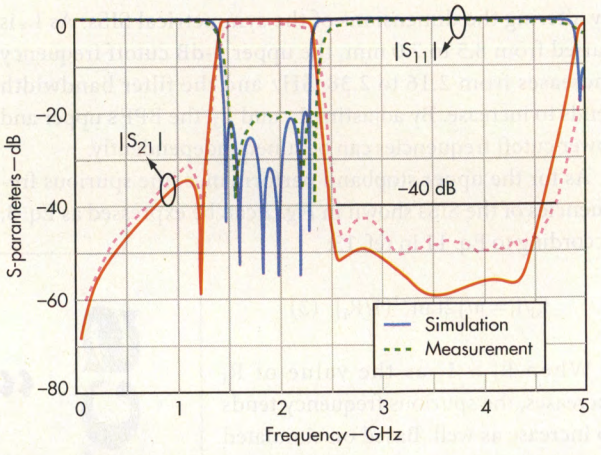
4. This BPF is fabricated with the stepped-impedance stub-loaded resonator design approach.



ter frequency of 1.88 GHz and fractional bandwidth of 48%. The minimum passband insertion loss was measured at about 0.6 dB. The passband return loss is better than 20 dB.

The BPF has two transmission zeros located at 1.23 and 2.61 GHz, with 41.2 and 47.9 dB rejection, respectively, for high passband selectivity. The filter also achieves better than 32-dB attenuation in the lower stopband while the measured upper stopband extends to 4.7 GHz (namely $2.5f_0$) with better than 40-dB attenuation.

By making use of a design approach with a stepped-impedance stub-loaded resonator, this BPF achieves compact size, high selectivity, and good out-of-band rejection. Its upper and lower cutoff frequencies can be tuned independently, with measured performance showing a compact BPF with 48% bandwidth. mw



5. The results of measurements and predictions from electromagnetic (EM) simulation software on the novel BPF agree fairly closely.

ACKNOWLEDGMENTS

The authors wish to thank all readers for their appreciation of this work. The PCB substrate used in this work is TLX-8 from Taconic Plastics (www.4taconic.com). The CAE simulations carried out in this report were based on the use of Advanced Design System (ADS) simulation tools from Agilent Technologies.

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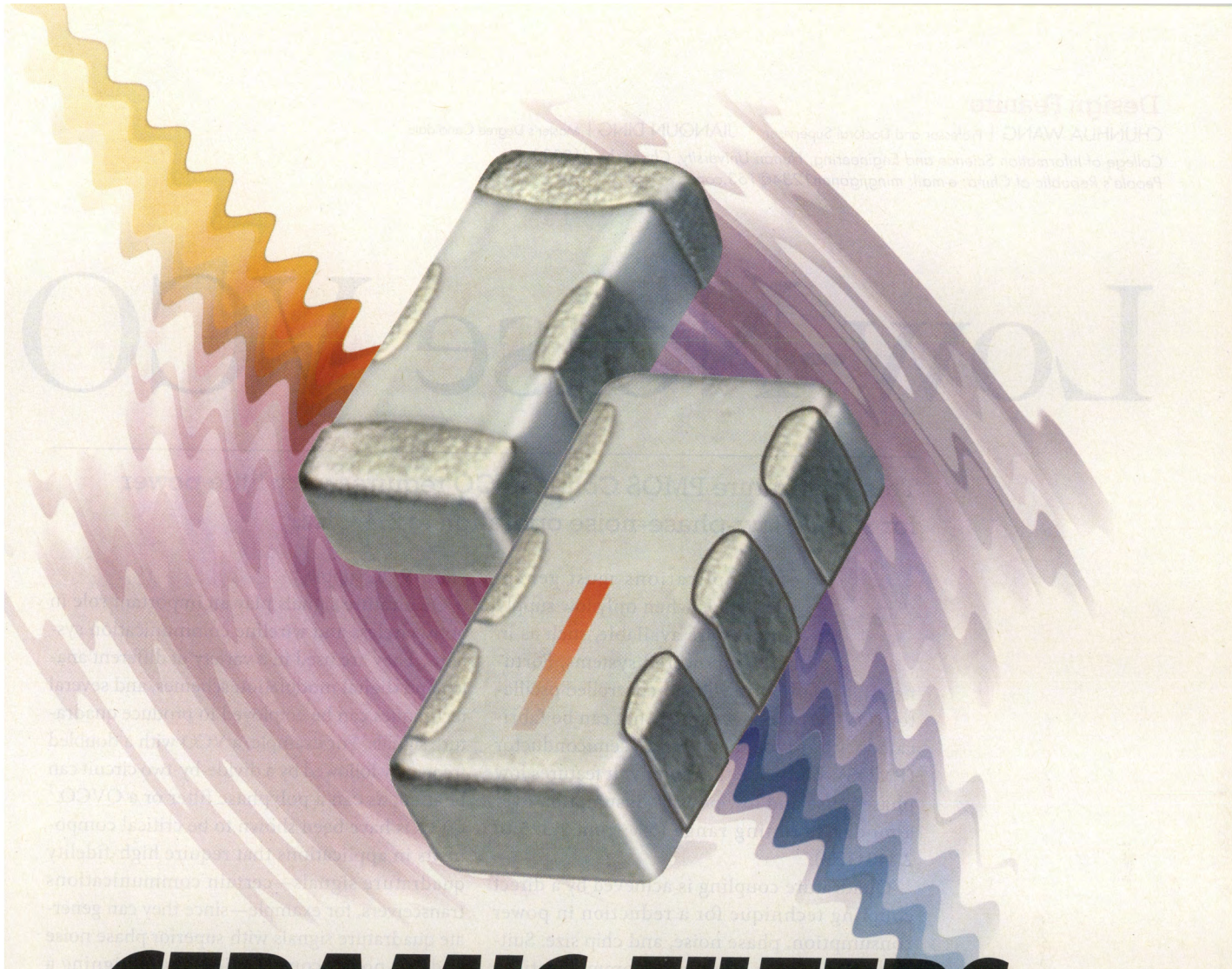

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Low-Noise VCO

This quadrature PMOS Colpitts VCO requires very little power for stable, low-phase-noise operation at 2.4 GHz.

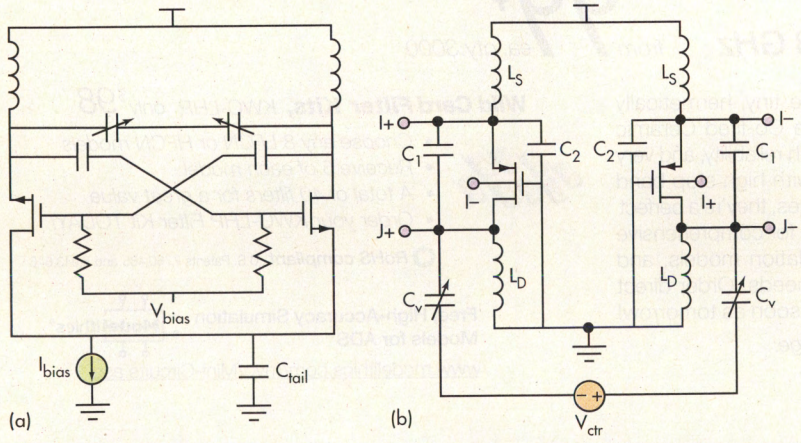
Some applications must generate signals when only low supply voltages are available, such as in battery-powered systems. Fortunately, a quadrature voltage-controlled oscillator (QVCO) was developed that can be fabricated with p-type metal-oxide-semiconductor (PMOS) solid-state technology. It features low phase noise and low power consumption over a moderate tuning range (5% from 2.325 to 2.440 GHz).

Quadrature coupling is achieved by a direct coupling technique for a reduction in power consumption, phase noise, and chip size. Suitable for a variety of wireless communications applications, the Colpitts-topology QVCO adds capacitive feedback to achieve an enhanced voltage swing over its tuning range. The prototype QVCO boasts phase noise of −128.2 dBc/Hz offset 1 MHz from a 2.34-GHz carrier while consuming only 0.535 mW power from a +0.55-VDC supply.

Quadrature signals play an important role in both wireless and wireline communication systems. They are used in a variety of different analog and digital modulation schemes, and several techniques can be employed to produce quadrature signals. For example, a VCO with a doubled frequency followed by a divide-by-two circuit can be used, as can a polyphase filter or a QVCO.[1] QVCOs have been shown to be critical components in applications that require high-fidelity quadrature signals—certain communications transceivers, for example—since they can generate quadrature signals with superior phase noise and low power consumption. But designing a source capable of high-spectral-purity quadrature signals with low power consumption is still a challenge.

The simplest and most convenient way to implement a QVCO is by coupling two symmetric inductive-capacitive (LC) tank VCOs.[2-4] The main drawback to this approach is the typically poor phase noise that results. In 2008, a Class-C VCO topology was proposed with a theoretical phase-noise improvement of 3.9 dB for the same current consumption as a standard differential-pair LC tank oscillator.[5] Following that approach, several new Class-C VCO topologies have been proposed.[5-10] However, all of these designs are based on a single differential pair configuration, similar to a traditional VCO topology.[5]

In quest of quadrature signals with low noise at low power levels, a new oscillator configuration was explored. In contrast to VCOs with traditional cross-coupled differential-pair LC-tank topologies, this new approach is a PMOS Colpitts QVCO. By employing a direct bulk



1. These block diagrams present two different VCO architectures: (a) the traditional cross-coupled, differential-pair Class-C VCO and (b) the new PMOS Colpitts QVCO.

Runs on Low Voltage

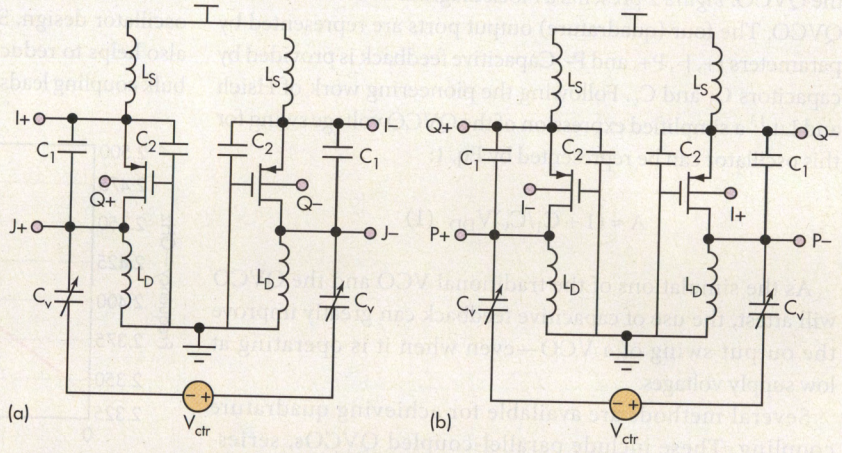
coupling technique, this novel QVCO topology enables quadrature coupling without requiring additional transistors. In addition, it provides an enhanced voltage swing under a low supply voltage by means of a capacitive feedback technique.

Figure 1(a) shows a cross-coupled, differential-pair, Class-C oscillator. Bias current is inserted between the NMOS source and ground and, in most cases, this is accomplished by means of a tail transistor. To filter high-frequency noise from the bias current, a large tail capacitance is employed; together with an resistive-capacitive (RC) bias network, this can prevent the transistors from working in the deep triode region.[5]

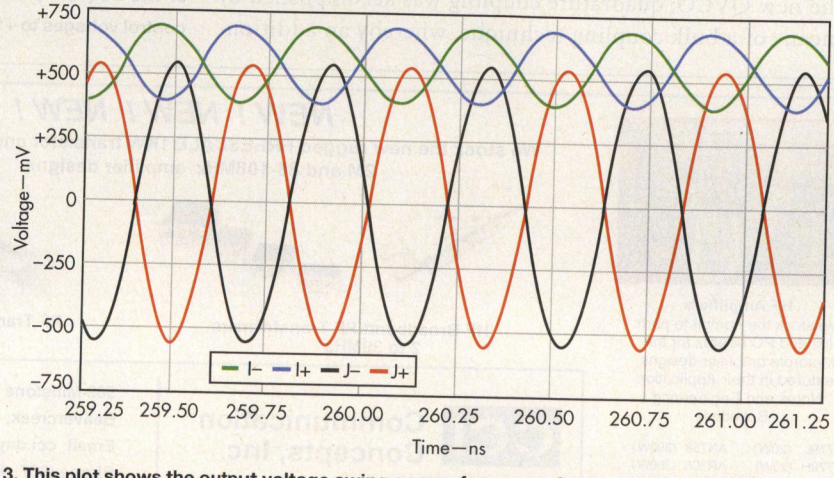
Figure 1(b) shows the new PMOS Colpitts VCO architecture. For a given amount of current, NMOS transistors are smaller than PMOS transistors for the same value of transconductance. However, PMOS transistors exhibit about 10-dB less 1/f noise compared with NMOS transistors. PMOS transistors also feature lower hot-carrier-induced white noise than NMOS transistors. As a result, PMOS-based VCOs can achieve better phase-noise performance than NMOS-based VCOs.[11] Based on these considerations, a new VCO core was designed using a full PMOS transistor architecture.

Rather than using transistors (and their noise contributions) for quadrature signal coupling, direct bulk coupling is employed in this new QVCO to improve phase-noise performance while reducing power consumption and semiconductor chip area. To improve the oscillator's voltage swing while operating with a low supply voltage, capacitive feedback is employed. The typical source of bias current has been replaced by an on-chip inductor, L_D, leaving maximum voltage headroom for the oscillating signal and dramatically reducing the phase noise.[12] This technique is especially useful in low-supply-voltage topologies.

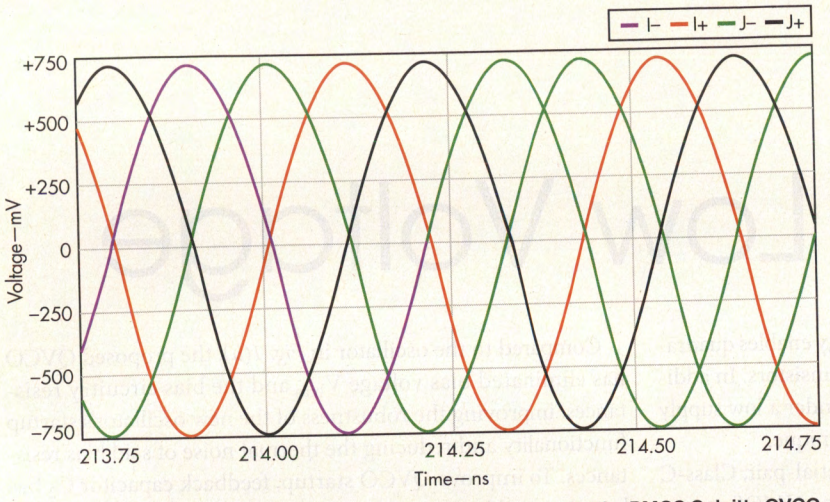
Compared to the oscillator in *Fig. 1(a)*, the proposed QVCO has eliminated bias voltage V_{bias} and the bias circuitry resistances, improving the robustness of the new oscillator's startup functionality and reducing the thermal noise of said bias resistances. To improve QVCO startup, feedback capacitor C_2 has been inserted between the PMOS source and gate. The greatest difference between the new QVCO topology of *Fig. 1(b)* and the traditional topology of *Fig. 1(a)* is that the new design employs a Colpitts architecture instead of the traditional cross-coupled differential-pair LC tank topology.



2. This is a more detailed schematic diagram of the proposed PMOS Colpitts QVCO.



3. This plot shows the output voltage swing across frequency for the traditional VCO.



4. This plot shows the output voltage swing across frequency for the PMOS Colpitts QVCO.

The greatest difference between the proposed QVCO and the traditional cross-coupled differential-pair LC-tank configuration of *Fig. 1(a)* is the use of the Colpitts architecture in the QVCO. *Figure 2* presents a block diagram of the new PMOS QVCO. The four (quadrature) output ports are represented by parameters J+, J-, P+, and P-. Capacitive feedback is provided by capacitors C_1 and C_v. Following the pioneering work of Hsieh and Lu,[13] a simplified expression of the QVCO voltage swing for this oscillator can be represented by Eq. 1:

$$A \approx (1 + C_1/C_v)V_{DD} \quad (1)$$

As the simulations of the traditional VCO and the QVCO will attest, the use of capacitive feedback can greatly improve the output swing of a VCO—even when it is operating at low supply voltages.

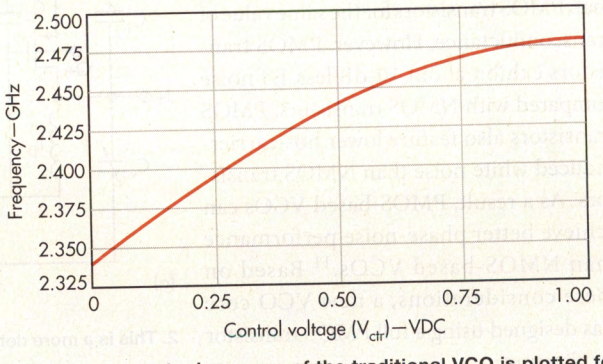
Several methods are available for achieving quadrature coupling. These include parallel-coupled QVCOs, series-coupled QVCOs, and back-gate-coupled QVCOs.[14-16] For the new QVCO, quadrature coupling was accomplished by means of a bulk coupling technique, whereby an additional transistor or capacitor was not needed to provide the coupling. The use of direct bulk coupling can reduce the supply-voltage requirements.

For a MOSFET device, the threshold voltage is governed by the body effect[17]:

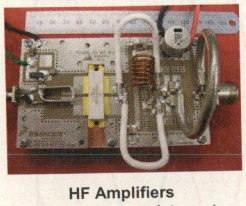
$$V_t = V_{t0} + \gamma[(|2\Phi_F + V_{SB}|)^{0.5} (|2\Phi_F)|)^{0.5} \quad (2)$$

where:
V_{SB} = the sideband voltage;
V_{t0} = the threshold voltage for $V_{SB} = 0$ V;
γ = the bulk effect parameter, with a typical value between 0.3 and 0.4 $V^{0.5}$; and
Φ_F = a physical parameter with typical value of 0.3 V.

From Eq. 2, it is known that by applying both DC and AC bulk coupling, the effective threshold voltage can be reduced; this is desirable for achieving a low-supply-voltage, low-power oscillator design. Since forward biasing of the bulk coupling also helps to reduce flicker noise in a PMOS VCO, the use of bulk coupling leads to a coupling factor of less than unity.[18] As a



5. The frequency tuning range of the traditional VCO is plotted for control voltages to +1 VDC.



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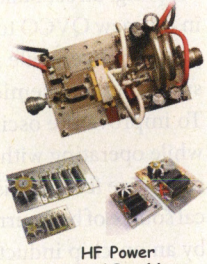

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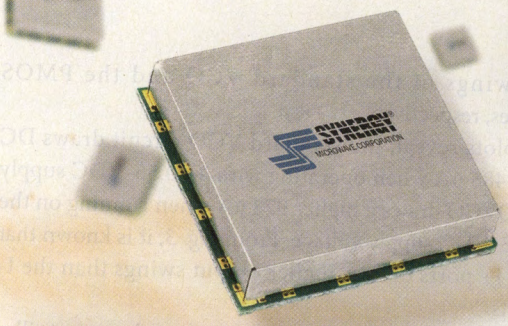

Model	Frequency (MHz)	Tuning Voltage (VDC)	DC Bias VDC @ I [Max.]	Phase Noise @ 10 kHz (dBc/Hz) [Typ.]
HFSO640-5	640	0.5 - 12	+5 @ 35 mA	-151
HFSO745R84-5	745.84	0.5 - 12	+5 @ 35 mA	-147
HFSO776R82-5	776.82	0.5 - 12	+5 @ 35 mA	-146
HFSO800-5	800	0.5 - 12	+5 @ 30 mA	-146
HFSO914R8-5	914.8	0.5 - 12	+5 @ 35 mA	-139
HFSO1000-5	1000	0.5 - 12	+5 @ 35 mA	-141
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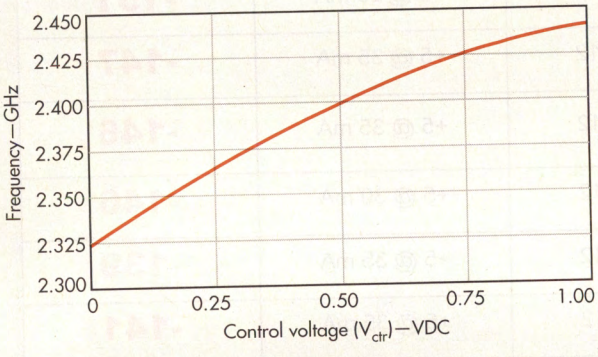
"By employing a direct bulk coupling technique, this novel QVCO topology enables quadrature coupling without requiring additional transistors."

result, the direct bulk coupling technique can also improve the oscillator's phase-noise performance. The design parameters for the QVCO are as follows: C_1 = 429.4 fF; C_2 = 664.9 fF; L_S = L_D = 7.1 nH; and dimensional parameters of 120/0.18 μm for M_1, M_2, M_3, and M_4.

By way of comparison, circuits for the traditional VCO and the proposed PMOS Colpitts QVCO were fabricated in monolithic form using 0.18-μm RF CMOS semiconductor technology from Taiwan Semiconductor Manufacturing Company Ltd. (TSMC; www.tsmc.com). The two oscillator circuits were compared in terms of output voltage swing, frequency tuning range, and phase noise. *Figures 3 and 4* show the output voltage swings of the standard VCO and the PMOS QVCO cores, respectively.



6. The frequency tuning range of the new Colpitts QVCO is plotted for control voltages to +1 VDC.

As the plots reveal, the standard VCO circuit draws DC current of 486 μA when operating with a +0.55-VDC supply while the QVCO draws a higher 972 μA when running on the same +0.55-VDC supply voltage. From *Fig. 3*, it is known that the J- and J+ ports deliver higher output swings than the I- and I+ ports.

Figures 5 and 6 display the tuning ranges of the two oscillator architectures. The frequency range of the standard VCO is 2.34 to 2.48 GHz when control voltage V_{ctr} sweeps from 0 to +1 VDC; the tuning range of the QVCO is 2.325 to 2.440 GHz when working with the same V_{ctr} range. *Figures 7 and 8* offer phase-noise performance levels for the two oscillator architectures.

As these plots show, the best phase noise for the traditional VCO is −120.3 dBc/Hz offset far (1 MHz) from the carrier while the phase noise for the proposed QVCO is as low as −128.2 dBc/Hz offset 1 MHz from the carrier. In general, the performance of the new QVCO is superior to that of the traditional VCO, except for power consumption.

A figure of merit (FOM) was employed to evaluate the three performance parameters of frequency, phase noise, and power dissipation for the QVCO, as detailed in Eq. 3:

$$FOM = ELL(f_{offset}) - 20\log(f_0/f_{offset}) + 10\log(PDC/1\ mW) \quad (3)$$

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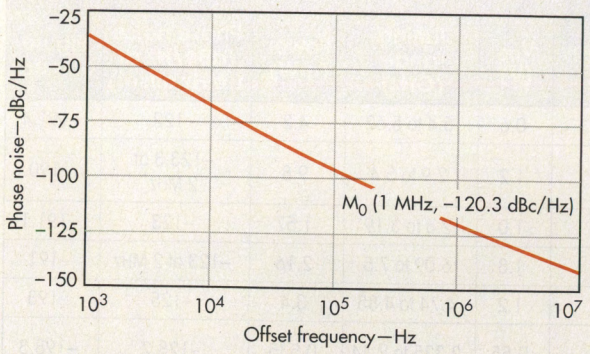
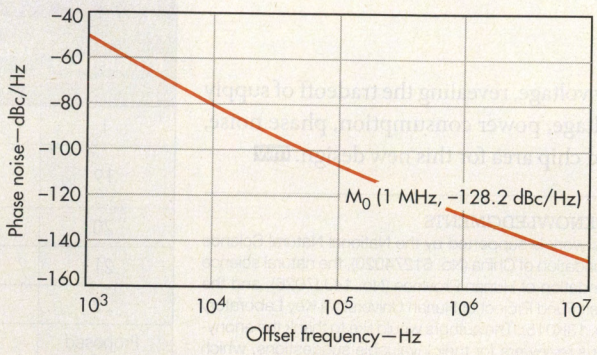


7. The phase noise of the traditional VCO is shown as a function of offset frequency.



8. The phase noise of the proposed Colpitts QVCO is shown as a function of offset frequency.

where:
PDC = the power dissipated by the oscillator core (in mW);
ELL(f_{offset} = the phase noise at the oscillator carrier frequency;
f_{offset} = the offset frequency from the oscillator carrier; and
f_0 = the center frequency.

The performance of the new QVCO is summarized in the table, and compared with recent work in QVCOs. The new low-voltage design exhibits the best FOM among these reported QVCOs, with higher FOM values indicating better oscillator performance.

In summary, a new PMOS Colpitts QVCO for a +0.55-VDC supply voltage has been developed and fabricated with a commercial 0.18-µm silicon CMOS technology. By using capacitive feedback and a direct bulk coupling technique, the quadratic oscillator achieves significant improvements in terms of power consumption and phase-noise performance, making it an attractive candidate for low-power RF/microwave applications requiring low supply voltages.

On the negative side, the proposed QVCO topology requires eight inductances, which leads to a larger required chip area than some other QVCO topologies. The four tail inductances, L_S, can be replaced by resistances or MOSFETs, but simulations show that this will degrade the phase-noise performance and power consumption. It will also require the use of a larger sup-

TUNING THROUGH VCO CHARACTERISTICS

VOLTAGE-CONTROLLED OSCILLATORS (VCOS) serve as the signal sources in many high-frequency systems. With frequency tuning controlled by an input voltage, they are quite versatile and useful in applications requiring variable frequency signals with different forms of phase and frequency modulation, as well as in frequency synthesizers operating with phase-locked loops (PLLs).

One of the simplest forms of VCOs can be created with a resonant inductive-capacitive (LC) circuit and an active device, such as a transistor, to provide signal gain. Tuning can then be achieved by changing the capacitance of the LC circuit by means of a device, such as a varactor diode, in which capacitance changes as a function of applied voltage. VCO packages and formats are often a function of application requirements, with tiny IC VCOs more often used in portable, low-power applications while larger, more rugged VCOs may be installed in professional and military radios and test equipment.

Specifying a VCO usually starts with its frequency tuning range and tuning voltage range, but commercial VCOs are typically sorted by a handful of specifications that allow straightforward comparisons among different products. Such parameters as output power, phase noise, spurious content, and power consumption are often used to compare different VCOs. Phase noise in itself can be difficult to compare because of how it is measured and characterized, at different offset frequencies from the carrier, and it is important when comparing VCOs to ensure phase-noise levels for different sources are compared at the same offset frequencies from the carrier.

Tuning linearity, which is a measure of how precisely and repeatedly the VCO changes frequency for a change in voltage, is an important parameter when tuning accuracy must be maintained in an application. Tuning sensitivity refers to the amount of voltage required for a given change in VCO frequency. Typically, higher tuning sensitivity is desirable for applications where less tuning voltage is available.

Amplitude flatness with frequency, which is a measure of how linear the VCO's output power remains across its tuning bandwidth, tends to be more difficult to maintain for VCOs with wider tuning bandwidths. The output flatness can also be impacted by other components.

Specifying a VCO can involve a number of different comparisons, such as tuning linearity, power consumption, and phase noise, and quite often achieving good performance in one parameter may mean some form of a tradeoff in another parameter. VCO suppliers are usually quite willing to discuss such tradeoffs.

ply voltage, revealing the tradeoff of supply voltage, power consumption, phase noise, and chip area for this new design. mw

ACKNOWLEDGMENTS

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COMPARING QVCO PERFORMANCE WITH RECENT RESULTS

Reference	Technology (nm)	V_{DD} (V)	Tuning range (GHz)	Power (mW)	Phase noise (dBc/Hz at 1 MHz offset)	Figure of merit
1	130	0.6	5.4 to 5.62	4.2	–122	–191.4
19	130	1.2	2.9 to 5.4	9.8	–123.3 at 2 MHz	–190
20	180	1.0	2.6 to 3.19	1.57	–123	–191.1
21	180	1.8	6.09 to 7.5	2.16	–123 at 2 MHz	–191
22	180	1.2	4.74 to 4.85	3.4	–125	–193
Proposed QVCO	180	0.55	2.325 to 2.44	0.535	–128.2	–198.3

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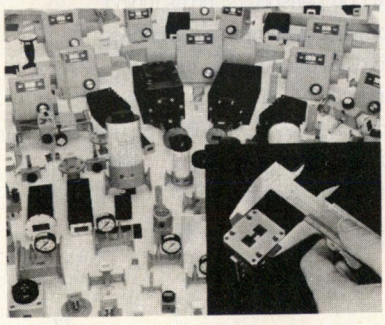

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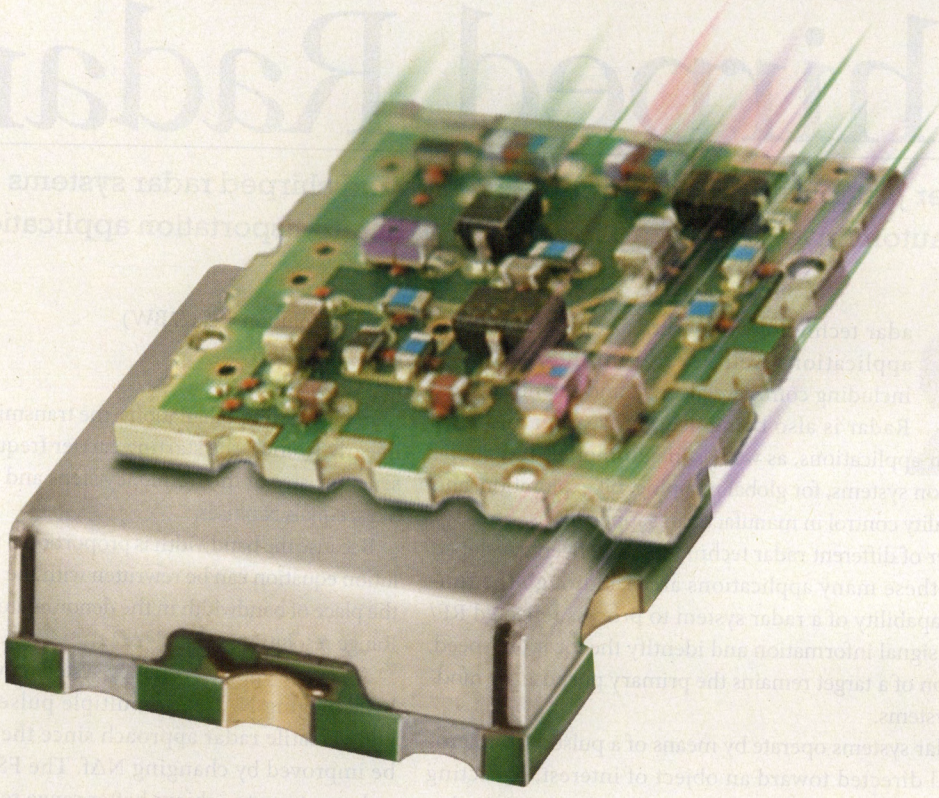

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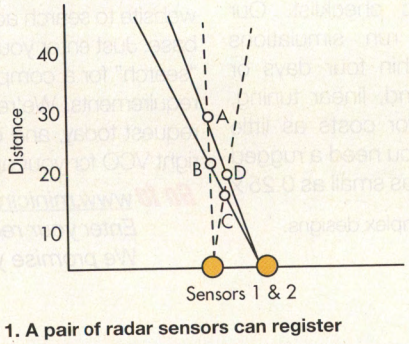
Radar technology now serves a broad range of applications well beyond military systems, including commercial, civil, and scientific uses. Radar is also being used in weather and law-enforcement applications, as well as in automotive and public transportation systems, for global mapping, oil exploration, and even for quality control in manufacturing applications.

A number of different radar techniques have been developed to support these many applications areas, although the fundamental capability of a radar system to process reflected RF/microwave signal information and identify the location, speed, and direction of a target remains the primary purpose for modern radar systems.

Most radar systems operate by means of a pulsed RF/microwave signal directed toward an object of interest, collecting reflected energy from the target using the same antenna that transmitted the pulses. However, some radar systems achieve this measurement using frequency-stepped chirped radar (FSCR) signals to transmit pulsed signals that conform to linear frequency ramps (known as chirps). FSCR radar systems are noteworthy for their capabilities to achieve excellent radar resolution using limited instantaneous bandwidth.

In such systems, successive pulses increase linearly in discrete frequency steps and pulse-modulated signals are transmitted via antennas. The echo signal that is reflected back to the antenna is combined with the transmitted signal to create a beat signal to calculate round trip time, τ, which is inversely proportional to the bandwidth (BW).

A radar system's range resolution is its capability to distinguish between two different targets at the same bearing with two different ranges.[1] Denoting c as the speed of light, range resolution can be found from the following relationship:



1. A pair of radar sensors can register different results for two different approaching targets.

$$\text{Range resolution} = c(1/2BW)$$

and

$$f_n = f_0 + n\Delta f$$

In performing calculations, the transmit frequency is stepped linearly from f_0, the starting carrier frequency, while Δf represents the size of the frequency steps and n = 1…N where each burst equals N pulses.

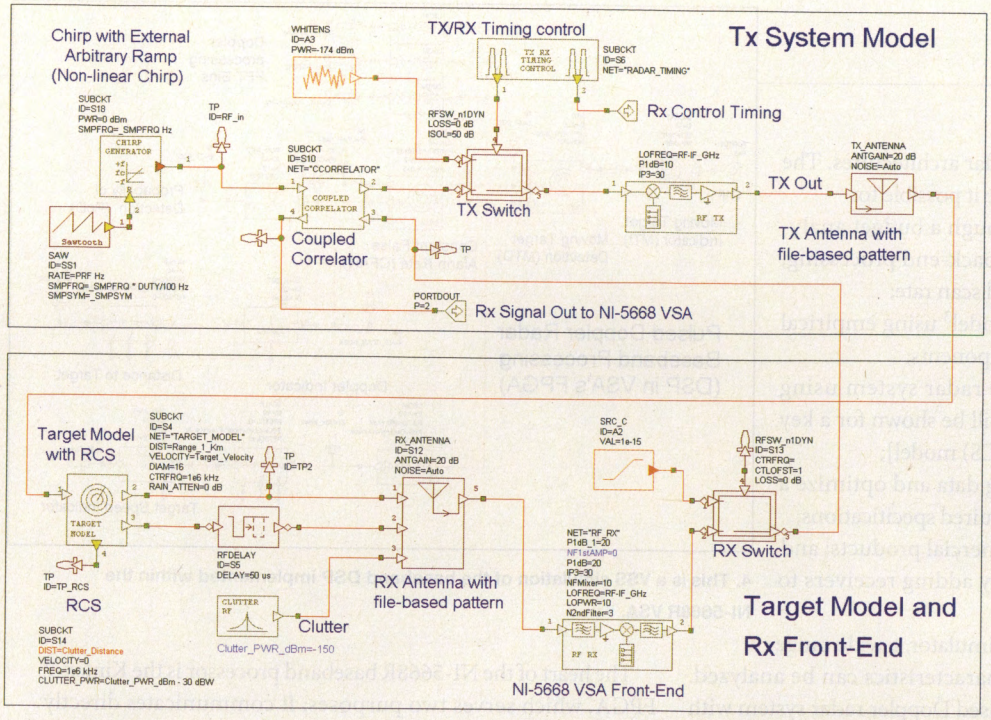
Because the bandwidth is proportional to $N\Delta f$, the range resolution equation can be rewritten with the frequency steps taking the place of bandwidth in the denominator:

$$\text{Range resolution} = c(1/2N\Delta f)$$

A linear stepped frequency system covers a wide bandwidth by achieving $N\Delta f$ with multiple pulses. This is a powerful and versatile radar approach since the range resolution can be improved by changing $N\Delta f$. The FSCR technique allows radar systems to achieve better range resolution with a design that implements a low instantaneous bandwidth and a relatively low sampling rate, ultimately lowering costs and reducing system complexity. Optimum design tools can yield a highly predictive design that facilitates hardware verification prior to amassing significant nonrecurring expenses and respin of custom microwave components.

The key functions of such a radar system can be modeled by means of modern computer-aided-engineering (CAE) systems simulation software using commercial off-the-shelf hardware as the radar system components. For example, by using high-frequency couplers, the same oscillator can serve as a stimulus for the mixer on the receiver as well as the transmitter. A commercial duplexer allows the radar system to transmit and receive in different time periods.

The radar receiver includes mixers, amplifiers, switches, isolators, phase detec-



2. Using VSS software, this is a model of a pulsed Doppler radar system with chirp capability.

tors, analog-to-digital-converter (ADC) circuitry, and signal processors. Accurate timing, phase, and frequency measurements between transmit and receive signals enable the radar system to determine target range and velocity.

Because the radar's target is mobile, the received signal will exhibit Doppler shifts. Differentiating between Doppler shift caused by clutter (objects that are not of interest but are returning reflected signals) and the true object of interest (the target) is a mathematically intensive problem. A moving target detection (MTD) caused by clutter can generate a false alarm. One way to mitigate this effect is to add additional sensors. A target's position can then be obtained by triangulating the bearing measurements from each pair of sensors.[2]

Figure 1 shows a system with two sensors where the triangulation measurement yields four results (A, B, C, and D) for the two actual targets which would result in a false alarm. The number of false detections will increase as more targets are added. Also, a chirped radar system can create its own Doppler coupling effects.

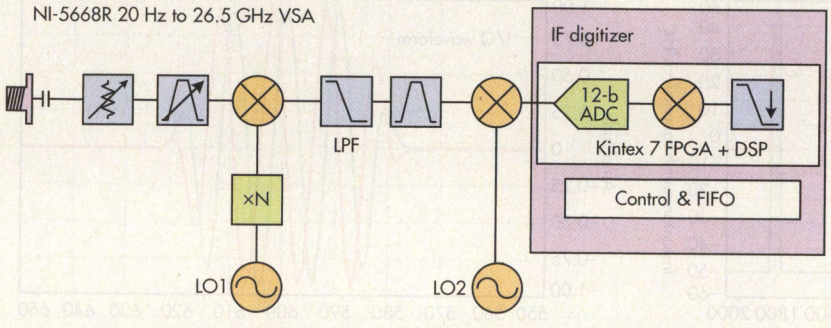
An effective radar system model can serve as an invaluable tool when evaluating different radar applications. As an example, the Visual System Simulator (VSS) software from AWR Corp. (www.awrcorp.com), a National Instruments (www.ni.com) company, features a number of useful tools built into the software, including a constant-false-alarm-rate (CFAR), MTD, and probability-of-detection (POD) indicators.

This type of signal processing is typically implemented by means of digital-signal-processing (DSP) circuitry within a receiver's baseband section (and following RF/microwave signal detection). To facilitate analysis, the VSS software works seamlessly with the firm's Microwave Office and Analog Office CAE software tools.

To build a representative model with such CAE software, the chirp signal as well as the quality of the receiver must be characterized to simulate real-world scenarios of multiple targets. By collecting echoes from chirped pulses and applying matched filtering and stochastic techniques, an image can be reconstructed with dimensional accuracy.

A host of computer-generated algorithms exist for radar signal processing to identify clutter and to reduce false alarms. Nevertheless, implementation of more-complex algorithms can be time consuming, especially when broadband coherent or correlated demodulation is required. Fortunately, FSCR systems can be modeled fairly easily using a broadband DSP-capable receiver.

To capture the frequency- and time-domain information for modeling radar signals, appropriate test equipment is required. A wideband vector signal analyzer (VSA), like the model NI-5668R from National Instruments, is one such example. This modular PXI-format analyzer, with available frequency ranges of 20 Hz to 14 GHz and 20 Hz to 26.5 GHz and as much as 765 MHz of instantaneous bandwidth, contains a powerful field-programmable gate array (FPGA) that can be used to simulate different radar detection algorithms.



3. This is the hardware, firmware, and software architecture of the model NI-5668 PXI vector signal analyzer (VSA).

Such advanced measurement hardware and modern CAE software tools provide the capabilities to dramatically improve the design efficiency, cost effectiveness,

and performance of advanced radar architectures. The hardware and software tools make it possible to:

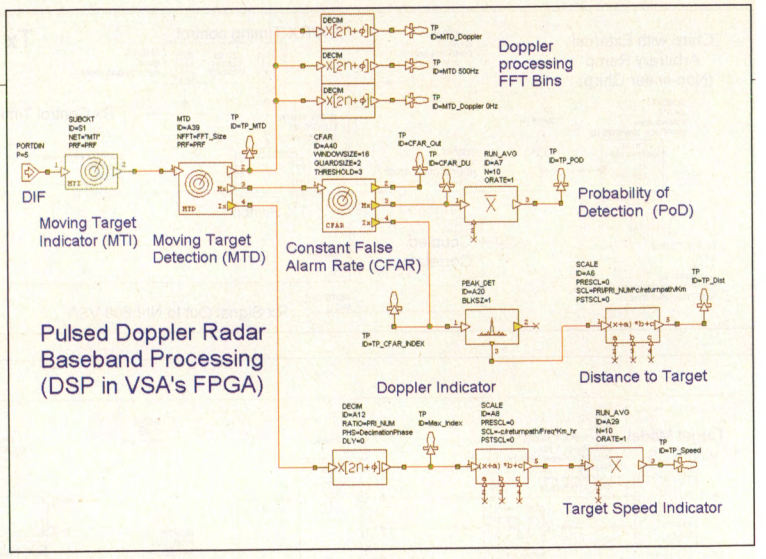
- Derive system constraints through a budget analysis which accounts for system back-end processing, dynamic range, noise figure, and scan rate;
- Build a pulse Doppler radar model[3] using empirical results and data files for subcomponents;
- Model subcomponents in the radar system using empirical results [an example will be shown for a key part of a radar-cross-section (RCS) model];
- Analyze system-level modeling data and optimize a radar system design to meet required specifications;
- Prototype a design using commercial products; and
- Extend the hardware system by adding receivers to simulate multiple sensors.

With a system-level software simulator, a wide range of radar-system performance characteristics can be analyzed. *Figure 2* shows the model of a pulsed Doppler radar system with chirp capability. With VSS, the components of a basic radar architecture can be modeled using system-level primitive files, including actual or predicted S-parameter data files.

Thus, a chirp generator can be specified using performance parameters while a radar antenna can be modeled using an actual radar-cross-section (RCS) pattern. After the chirped signal has been transmitted and reaches the target, the reflected signal is detected by the antenna and passed to the receiver.

Figure 3 shows the receiver block diagram for this analysis. The NI-5668R VSA incorporates a Kintex-7 FPGA from Xilinx (www.xilinx). It can be programmed to model the receiver's various matched filter responses to differentiate a signal of interest from clutter by applying advanced detection schemes.

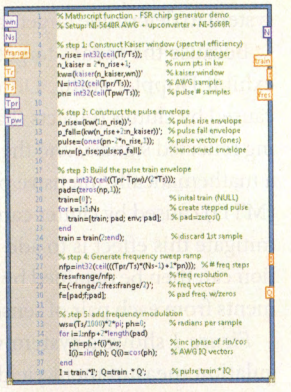
Figure 3 reveals the how various components within the NI-5668R VSR, such as a downconverter, software, firmware, and hardware, can be combined to model a basic radar system architecture. The final intermediate frequency (IF) in the system is digitized directly to avoid quadrature and offset errors common to zero-IF systems, thus avoiding calibration of in-phase/quadrature (I/Q) signal impairments.



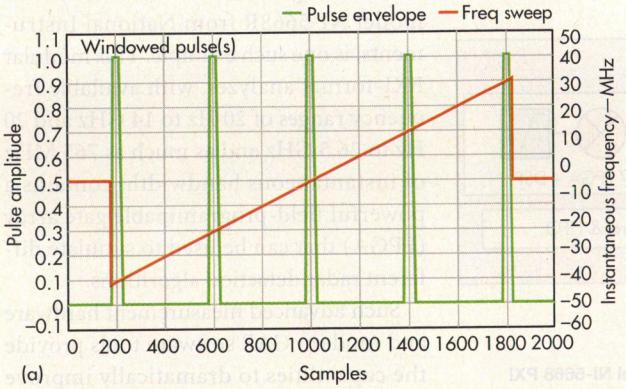
4. This is a VSS simulation of the baseband DSP implemented within the NI-5668R VSA.

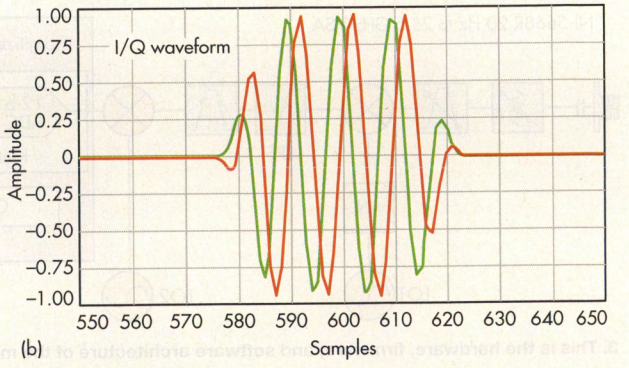
The heart of the NI-5668R baseband processor is the Kintex-7 FPGA, which serves two purposes: It communicates directly with the host program, processing and controlling the flow of data over the bus to the host computer, and it also performs complex digital-signal-processing (DSP) algorithms on digitized baseband signals. The VSA's instantaneous bandwidth of 765 MHz allows broadband microwave signals to be captured to 26.5 GHz by means of single data acquisitions.

Figure 4 shows a signal processing chain for radar detection. The computational algorithm for identifying targets and performing Doppler processing can be implemented in the FPGA. The notional model of the DSP section consists of an array of demodulators and threshold detectors. A fast Fourier transform (FFT) within the NI-5668R can be used to differentiate the Doppler components; however,

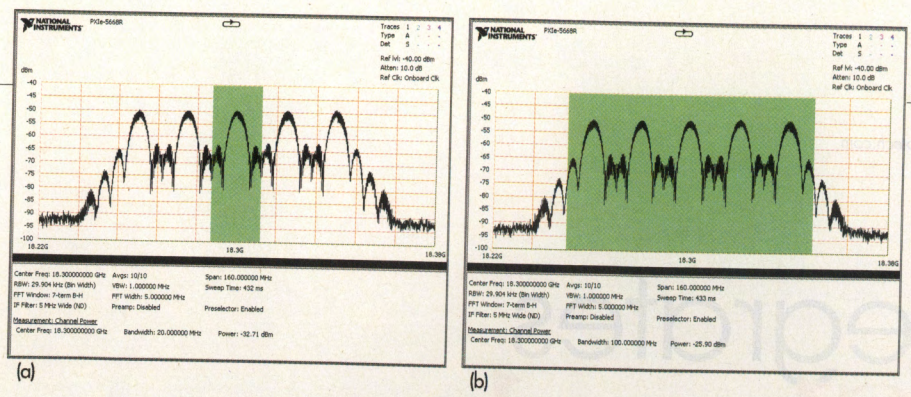


6. LabVIEW Mathscript code was used to verify the FSCR theoretical model and program the AWG.





5. This is the frequency-ramp and windowed pulse timing used to generate an FSCR I/Q waveform.



(a) (b)

7. These plots compare the spectral power in one frequency bin versus the total spectral power.

frequency spacing(s) could also be dynamic and uneven.

Implementation of FSCR requires precise synchronization with each frequency step along with coherent demodulation. In modern DSP-based receivers, increased bandwidth and coherent digital downconverters (DDCs) support broadband FSCR systems while maintaining dynamic range.

As shown in *Fig. 5*, a stepped chirp can be derived from a continuous chirp, but the pulses will gate the frequency ramp. In practice, the pulses can be modified with a window function, rather than using simple rectangular pulses. A Kaiser window is used in this example, and the data will show a more efficient spectral footprint. In terms of hardware, it is often easier to realize complex waveforms by means of an arbitrary waveform generator (AWG) that by other forms of signal generators.

In such a case, the pulses will gate the I/Q sweep generation of the AWG. The pulse shaping in the time domain will be due to the Kaiser window function. For the programming example of *Fig. 6*, LabVIEW Mathscript code from National Instruments was chosen because its syntax is familiar to DSP programmers. The input parameters "wn" and "Ns" correspond to the window-smoothing factor and the number of frequency steps.

The effectiveness of a model can be evaluated by comparison to measurements, an approach used to check the modeled spectrum of the FSCR signal. The modeled spectrum and the measured spectrum show nearly 15-dB suppression between frequency bins, with ultimate out-of-band rejection of at least 40 dB relative to any frequency peak and more than 66 dB relative to the total channel power of –26 dBm.

In theory, each of the five spectral components will share power equally, since the maximum spectral energy for each pulse is centered on its frequency at the time of the chirp. Parseval's Theorem states:

$$\sum_{n=1}^{NP} (pv_n^2) = \sum_{n=1}^{NB} \left(\int_{n-bw/2}^{n+bw/2} F\{p_n^2\} \right)$$

where:

$(pv^2)_n$ = the energy per pulse;

NP = the number of pulses; and

NB = the number of frequency bins (NP = NB).

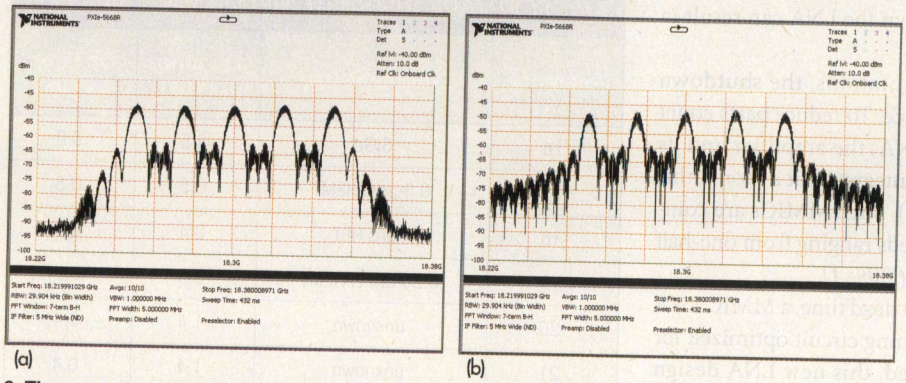
The integrated power in the frequency bins should equal the sum of the power of all the pulses. By extension, assuming each pulse has a unique frequency, each pulse's energy in the frequency domain will be equally weighted.

Figure 7 provides a comparison of the spectral power in one frequency bin to the total power. The measured power in the central 20-MHz portion of the FSCR spectrum is –32.71 dBm while the entire spectrum measures -25.9 dBm. The measured power ratio is:

$$P_{20}/P_{100} = 10\exp\{[-32.71 - (-25.9)]/10\} = 0.208 \approx 1/5$$

Figure 8 compares the measured spectral efficiency due to the Kaiser windowing of the pulse envelope. The windowed version shows nearly a 20 dB improvement in the adjacent channel.

Modern radar systems can be quite complex. FSCR systems in particular are signal processing intensive, so it is quite useful to model the hardware and algorithms comprising these systems. Fortunately, modern software and test tools make it possible to effectively model even complex radar systems before committing to any lineup of hardware components and system algorithms. mw



(a) (b)

8. These are the measured results of the Kaiser windowed signals (left) versus traditional rectangular pulses (right).

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Design Feature

CHIN LEONG LIM | Engineer
Avago Technologies, Penang, Malaysia, www.avagotech.com

LNA Integrates Fast Shutdown Function

This low-noise amplifier includes an on-board switching function with industry-leading switching speed, thus reducing dead time when a transceiver shifts between transmit and receive modes.

Time can be critical in modern communications systems, such as time-division-duplex (TDD) wireless systems. Although frequency-division-duplex (FDD) approaches have long dominated cellular communications, limited spectrum and high cost are making TDD approaches more attractive, as evidenced by growing use of standards such as time-division synchronous code-division-multiple-access (TD-CDMA) and TD Long Term Evolution (LTE).[1,2] To prevent dead time in TDD switching, reception should ideally commence as soon as transmission has ended.

However, a cellular system's low-noise amplifier (LNA), which is typically shut down during transmission to prevent damage of active devices and receiver overload, can suffer delays in returning to its active state. Because of these delays, a "guard time" of about 2% of the channel time must be allocated as settling time for the wireless system's hardware.[3] When transmission recommences, any delay in shutting down the LNA can result in temporary receiver overload.

In older wireless system implementations, the shutdown function was external to the LNA device. To reduce parts count and enable miniaturization of the LNA, the amplifier and its shutdown function are now typically integrated in a monolithic microwave integrated circuit (MMIC). Such MMICs are commercially available with switching speeds ranging from one-half microsecond to several microseconds *(Table 1)*.

To enable further reduction in TDD dead time, a MMIC LNA was developed with an on-chip switching circuit optimized for speed. In addition to switching speed, this new LNA design must also meet the stringent noise and linearity requirements demanded by the cellular infrastructure.

In designing this MMIC LNA with faster shutdown speed, the location of the switch location within the circuit can influence the switching speed and affect the ease of integration for the device. A shutdown function can be added to virtually any MMIC by connecting a switching transistor in series with the voltage supply, V_{dd} *(Fig. 1)*. The transistor's current-carrying capability must match the MMIC's current consumption, typically ranging from tens to hundreds of milliamperes. The single-supply MMIC design of *Fig. 1* generates its gate bias (V_{GS}) using voltage reference V_D.

Diode resistance R_D limits the current through diode D. The forward-biased diode generates wideband noise and capacitance C is required to suppress that noise. Unfortunately, this capacitance requires a finite amount of time to charge and discharge.

TABLE 1: COMPARING ON/OFF TIMES FOR VARIOUS SWITCHED MMIC LNAs

Reference	Technology	On time, t_{on} (µs)	Off time, t_{off} (µs)
16	SiGe	8.5	5.0
17	0.5-µm eHEMT	0.5	0.5
18	GaAs HBT	0.4	0.3
19	0.25-µm ePHEMT	0.6	unknown
20	unknown	1.4	0.4
21	unknown	1.4	0.4
Current	0.25-µm ePHEMT	0.05	0.02

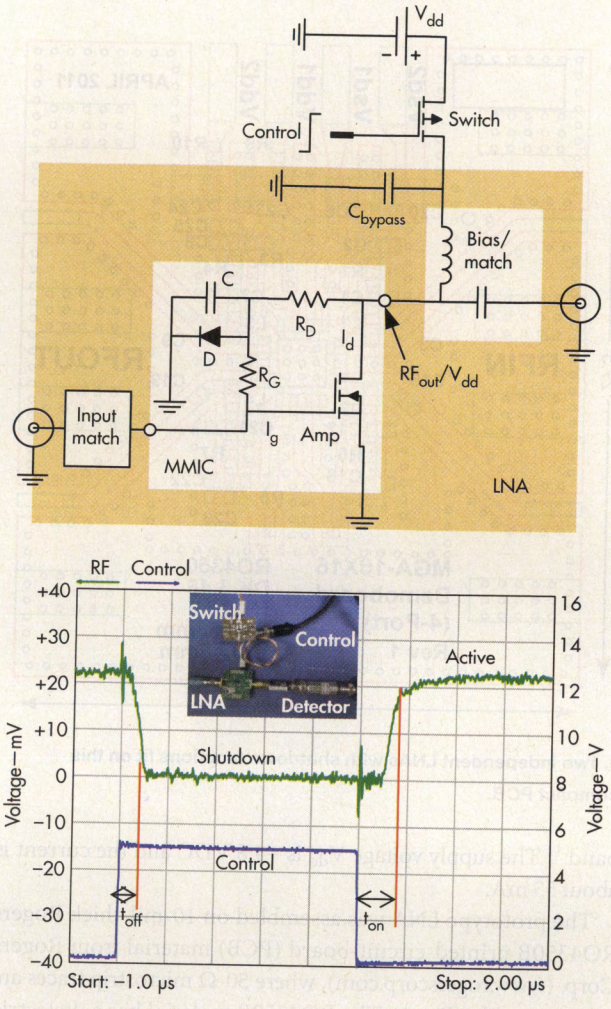
The time constant associated with R_DC slows the rise time of the gate voltage V_{GSat} with power on, while R_GC delays gate bias V_{GS} from decaying to zero at shutdown.

Additionally, the combination of the supply bypass capacitor C_{bypass} and the switching transistor's "on" resistance (typically 25 to 200 Ω[4]) can also slow down the switching speed, particularly if a large value of C_{bypass} is chosen to effectively suppress supply transients. Because of these unavoidable time constants, an externally switched MMIC LNA will be limited in switching speed. Typical switching speeds range from 1 to 4 µs *(Fig. 1)*.[5]

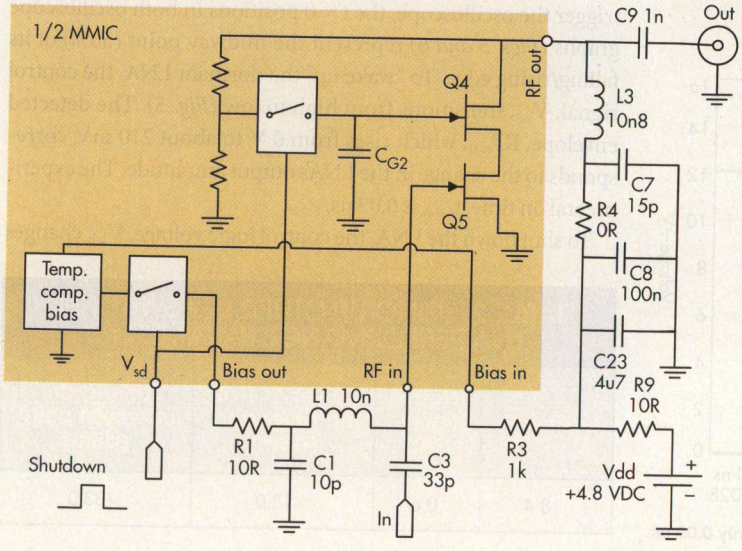
The MMIC is fabricated by means of a proprietary 0.25-µm enhancement-mode pseudomorphic high-electron-mobility-transistor (ePHEMT) process on 6-in. wafers.[6,7] The MMIC integrates dual amplifiers, adjustable active biasing, and shutdown functions housed in a 16-pin 4 × 4 × 0.85 mm quad-flat-no-lead (QFN) package.[8,9] A cascade configuration (Q4-5) was chosen for the amplifier because one common-source stage cannot meet the target gain at S-band *(Fig. 1)*.

The cascade configuration's bottom gate is biased by a temperature-compensated voltage reference and the top gate is biased by a resistor divider. During shutdown, both gates are disconnected from their respective biasing sources by internal transistor switches. Because gate current is in the microampere range, small switching transistors can be used for this purpose. Although interrupting either one of the two gate supplies is sufficient to shut down the cascade amplifier configuration, switching both gates simultaneously improves the forward isolation and the switching speed of the LNA switch function.

Because the switches are fabricated on the same high-speed process as the RF amplifier, their propagation delays are small in comparison to the overall switching time. On the other hand,



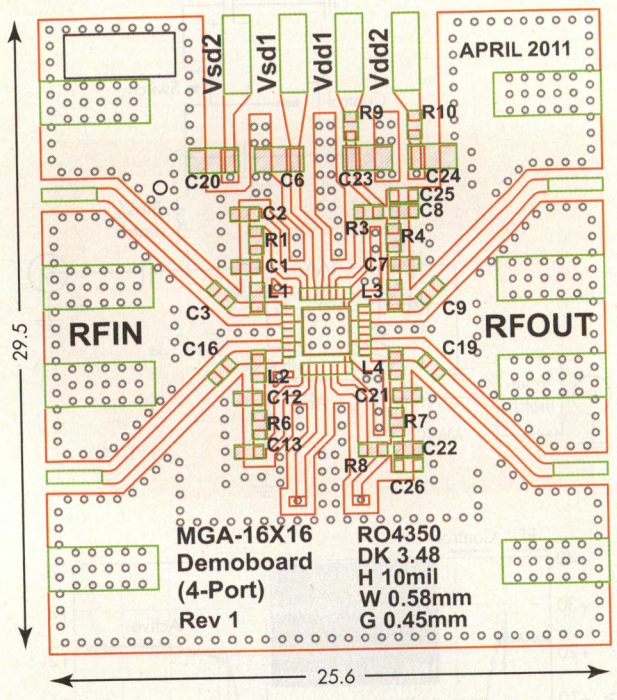
1. This shutdown function can be adapted to any LNA by inserting a switch into the supply line, but it will be limited to switching speed of no faster than 1 µs.



2. This monolithic integration of the LNA shutdown function reduces the parts count and the switching time.

the RC components, required to bias and decouple the cascode gates, can slow down the switching speed through their time constants. Since the biasing components are necessary, the only way to mitigate their effect on the switching speed is to choose the smallest usable values for capacitances C_1 and C_{G2} and the connected resistances.

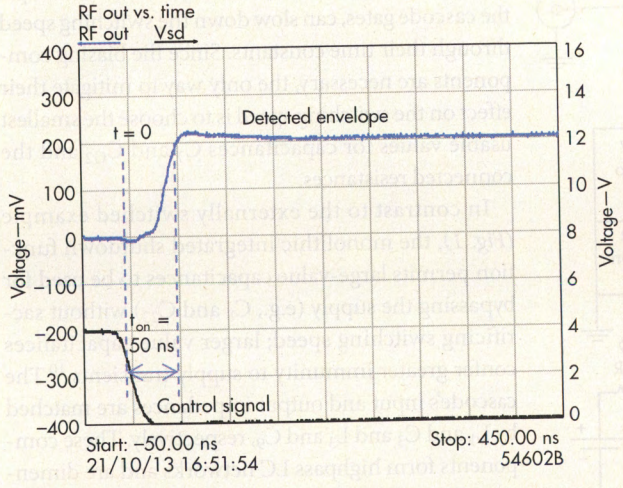
In contrast to the externally switched example *(Fig. 1)*, the monolithic integrated shutdown function permits large-value capacitances to be used for bypassing the supply (e.g., C_8 and C_{23}) without sacrificing switching speed; larger value capacitances confer greater immunity to supply transients.[10] The cascode's input and output impedances are matched by L_1 and C_3 and L_3 and C_9, respectively. These components form highpass LC networks and are dimensioned for operation at 2.6 GHz, i.e., the UMTS VII



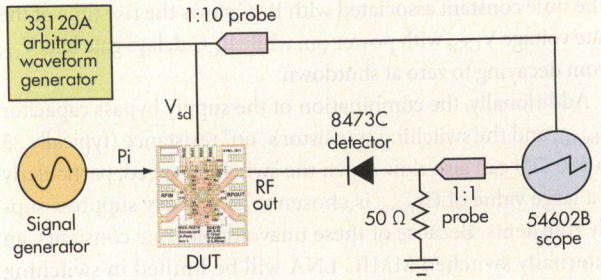
3. Two independent LNAs with shutdown functions fit on this compact PCB.

band.[11] The supply voltage V_{dd} is +4.8 VDC and the current is about 55 mA.

The prototype LNA was assembled on 10-mil-thick Rogers RO4350B printed-circuit-board (PCB) material from Rogers Corp. (www.rogerscorp.com), where 50-Ω microstrip traces are 0.58 mm wide *(Fig. 3)*. The RO4350B material has a dielectric constant of typically 3.48 in the z-direction at 10 GHz. An FR-4 backing layer provides rigidity and increases the stack height to 1.6 mm to suit edge-launch SMA receptacles.



5. The prototype switched LNA exhibited turn-on time of only 0.05 µs.



4. This simple block diagram shows the test setup for measuring the LNA's switching speed.

MAKING MEASUREMENTS

The time taken to switch from one LNA state to another has been variously referred to as "switching time/speed/rate" and "power settling time" due to absence of a standardized terminology. The turn-on time, t_{ON}, is measured from 50% of the control signal to 90% of the final output amplitude.[12,13] The turn-off time, t_{OFF}, is similarly defined. The setup *(Fig. 4)* for evaluating the LNA switching speed follows[13] An RF signal generator feeds the LNA with a 2.6 GHz, 0-dBm carrier.

A pulse generator provides the logic-level control signal for the switches. The envelope of the amplified carrier is detected by a low-barrier Schottky diode detector. The output of the detector is loaded with a 50-Ω feedthrough resistor to quicken its response time to between 8 and 12 ns.[14]

As part of the switched-LNA test setup, an oscilloscope displays the detected envelope and the control signal; the rising/falling edge of the control signal triggers the oscilloscope. Although a spectrum analyzer can replace the combination of a diode detector and an oscilloscope, the latter was selected for its fast response time compared to a spectrum analyzer.[15]

Measurements of the prototype switched LNA show it to exhibit the fastest turn-on and turn-off times in the industry *(Table 1)*.[16-21] Since the LNA's switch control signal was used to trigger the oscilloscope, the t = 0 positions in both oscilloscope graphs *(Figs. 5 and 6)* represent the midway point (50%) of its falling/rising edge. To "wake up" the dormant LNA, the control signal, V_{sd}, transitions from high to low *(Fig. 5)*. The detected envelope, RF_{out}, which rises from 0 V to about 210 mV, corresponds to the change in the LNA's output amplitude. The experimental on time, t_{ON}, is 0.05 µs.

To shut down the LNA, the control logic voltage, V_{sd}, changes

TABLE 2: SIZING UP OTHER KEY SWITCHED LNA PERFORMANCE PARAMETERS

Gain (dB)	Noise figure (dB)	Power at 1-dB compression (dBm)	Third-order intercept point (dBm)
18.4	0.6	+18.0	+33.0

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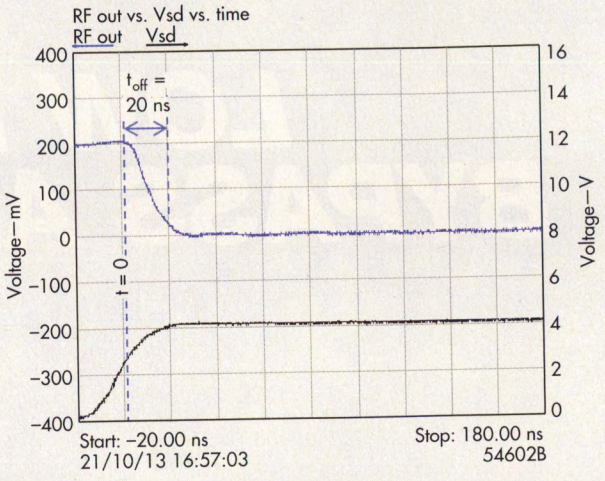
level from "LOW" to "HIGH" *(Fig. 6)*. Correspondingly, the detected envelope, RF_{out}, decreases from about 200 mV to 0 V. Referenced to the control-signal midpoint, the detected envelope (RF_{out}) requires 0.02 µs to drop to 10% of its maximum value.

The actual t_{OFF} time is most likely faster than the experimental value of 0.02 µs because it is very close to the diode detector's specified response time (≤0.012 µs). In addition to speed, the prototype switched LNA exhibited excellent performance levels in various other parameters, as detailed in *Table 2*.

In conclusion, this integration of the shutdown function with the LNA circuitry avoids the high current levels and slower speeds associated with external switching solutions. A combination of fast pHEMT technology, dual switches, and optimum circuit connections enable this design to be more than one order magnitude faster than its closest competitor. mw



6. The prototype switched LNA demonstrated turn-off time of only 0.02 µs.

ACKNOWLEDGMENTS

The author would like to thank H.A. Zulfa and Y.H. Chow for the IC design; S. Punithevati and M.D. Suhaiza for assembling the prototypes; and S.A. Asrul and the management of Avago Technologies for approving the publication of this work.

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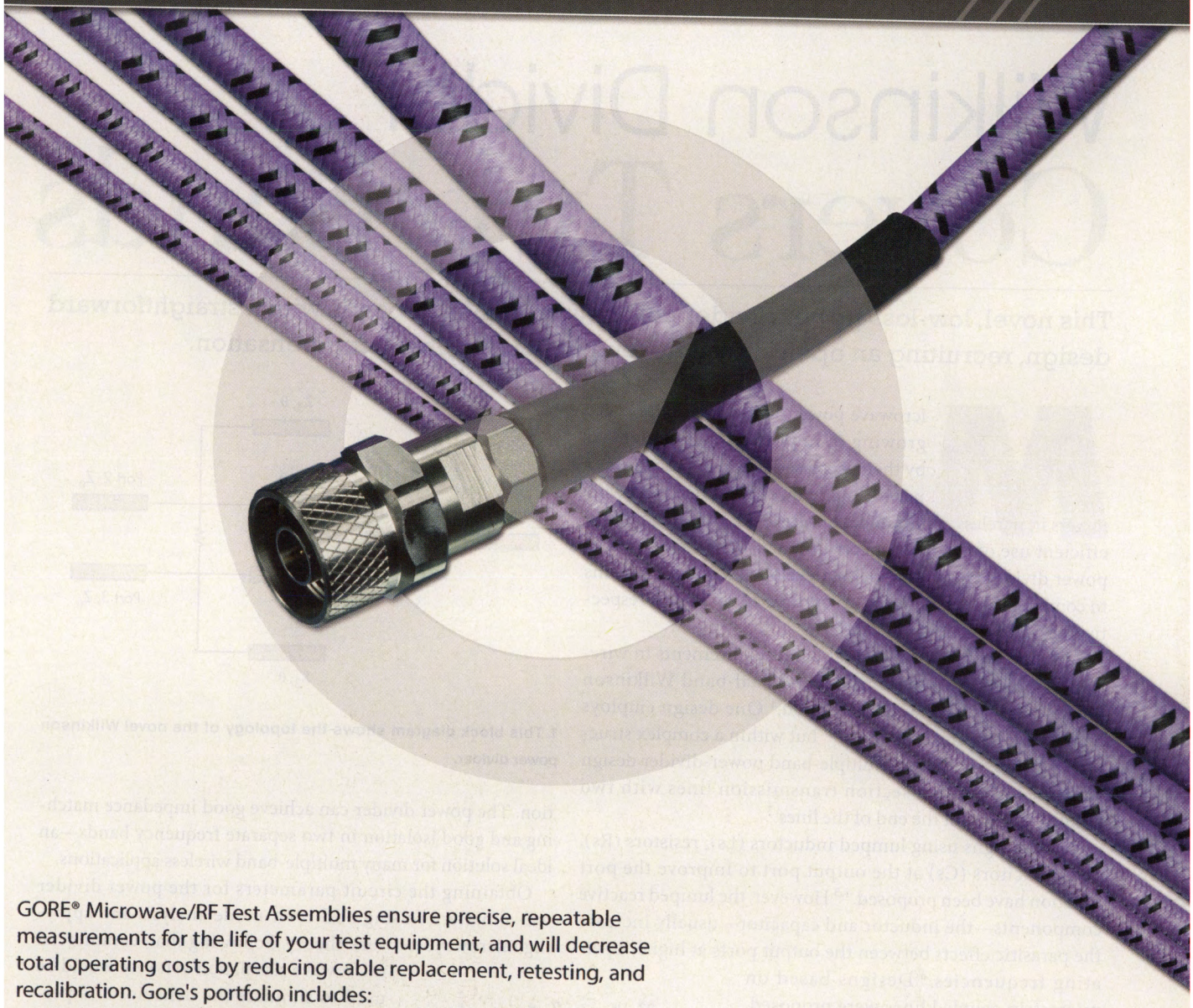

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Wilkinson Divider Covers Two Bands

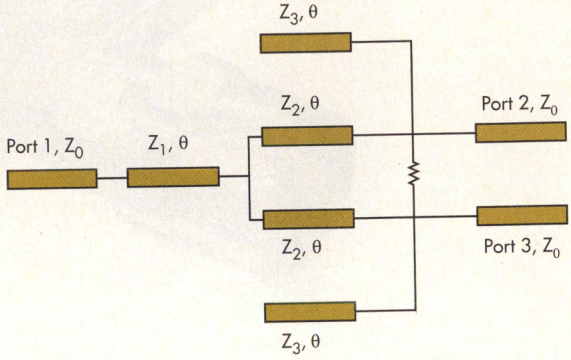
This novel, low-loss power divider covers two frequency bands with a straightforward design, recruiting an open stub at the output port for phase compensation.

Microwave power dividers/combiners are growing in importance, as evidenced by their increased use in antenna feeder networks, in power amplifiers, and with mixers in wireless communications systems. For effective and efficient use of available spectrum, dual-band and tri-band power divider circuits are being developed for these systems to combine the functions of two or three components, respectively, within one unit.

With the proliferation of dual-band requirements in wireless communication systems, several dual-band Wilkinson power dividers have been developed.[1] One design employs asymmetric transmission lines,[1,2] but within a complex structure. The most common multiple-band power-divider design is formed using two-section transmission lines with two lumped resistors at the end of the lines.[3]

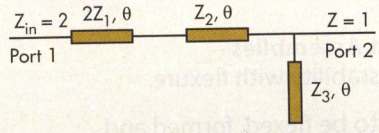
Some designs using lumped inductors (Ls), resistors (Rs), and capacitors (Cs) at the output port to improve the port isolation have been proposed.[4,5] However, the lumped reactive components—the inductor and capacitor—usually increase the parasitic effects between the output ports at higher operating frequencies.[3] Designs based on microstrip coupled lines were proposed, but each of these designs suffered some shortcomings.[6,7] To overcome the shortcomings of these previous dual-band power dividers, a new type of dual-band Wilkinson power divider with open stub at the output port was developed to aid the many expanding wireless communications applications.

Figure 1 shows the topology of the proposed Wilkinson power divider. The characteristic impedances of the three transmission lines are denoted as Z_1, Z_2, and Z_3; the lengths of the transmission lines are represented by L_1, L_2, and L_3, respectively. Transmission line Z_3 is used for phase compensation. The power divider can achieve good impedance matching and good isolation in two separate frequency bands—an ideal solution for many multiple-band wireless applications.



1. This block diagram shows the topology of the novel Wilkinson power divider.

Obtaining the circuit parameters for the power divider was a matter of using odd-even-mode analysis and applying microwave network theory. Following that, commercial microwave simulation software would be used to simulate the behavior and performance of the circuit, and a physical example of the circuit design would be made for measurement purposes.



2. This simple equivalent circuit of the power divider was used for even-mode analysis.

Because the power-divider circuit is symmetrical with respect to the middle line between the two output ports, there is no current flowing through the lumped components between the output ports in the even-mode situation. As a result, it is possible to produce the equivalent circuit shown in *Fig. 2* for the microwave power divider. To facilitate the analysis, all impedance values were normalized to 50 Ω of the port impedance.

By applying microwave network theory, the equality for the

ABCD matrix is given by Eq. 1:

$$\begin{bmatrix} \cos\theta & 2jZ_1\sin\theta \\ j\frac{\sin\theta}{2Z_1} & \cos\theta \end{bmatrix}\begin{bmatrix} \cos\theta & jZ_2\sin\theta \\ j\frac{\sin\theta}{Z_2} & \cos\theta \end{bmatrix}\begin{bmatrix} 1 & 0 \\ j\frac{\tan\theta}{Z_3} & 0 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \quad (1)$$

According to the relationship between the change in impedance and the ABCD matrix, in order to impedance match ports 1 and 2 of the power divider design, Eq. 2 must be satisfied for the input impedance, Z_{in}:

$$Z_{in} = (AZ + B)/(CZ + D) = 2Z \quad (2)$$

By solving Eqs. 1 and 2, Eqs. 3 and 4 can be obtained:

$$(Z^2{}_2Z_3 - 2Z^2{}_1Z_3 - 2Z^2{}_1Z_2 - Z^1Z_2{}^2)\tan_2\theta - Z_1Z_2Z_3 = 0 \quad (3)$$

$$Z_2Z_3 + 2Z_1Z_3 + 2Z_1Z_2 - Z_2{}^2\tan^2\theta - 2Z^2{}_1Z_2Z_3 - Z_1Z^2{}_2Z_3 = 0 \quad (4)$$

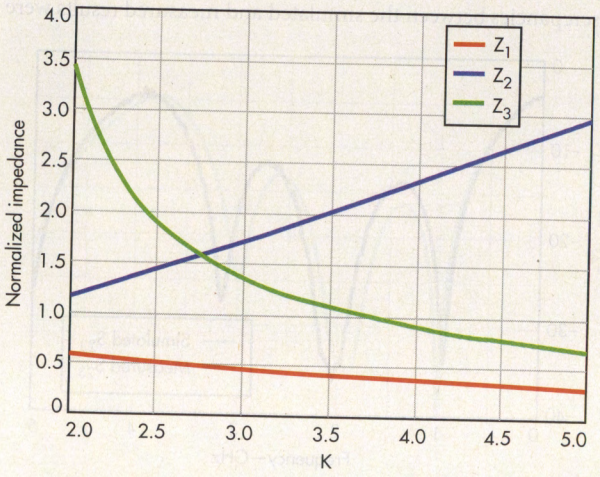
Figure 3 offers an equivalent circuit for the odd-mode analysis. In the odd-mode situation, there is a virtual ground in the middle of the two output ports, and impedance Z_1 doesn't work in this mode. Without the open stub, it is impossible to achieve impedance matching with this configuration.

From *Fig. 3*, the impedances of Z_A and Z_B can be written as:

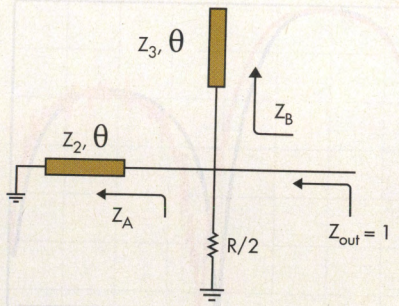
$$Z_A = j\tan\theta \quad (5)$$

$$Z_B = -j(Z_3/\tan\theta \quad (6)$$

To achieve impedance matching, impedances Z_A and Z_B



4. These curves trace the line impedances Z_1, Z_2, and Z_3 as functions of frequency.



3. This more elaborate model of the power divider circuit was used for odd-mode analysis.

must satisfy Eq. 7:

$$1/Z_A + 1/Z_B + 2/R = 1 \quad (7)$$

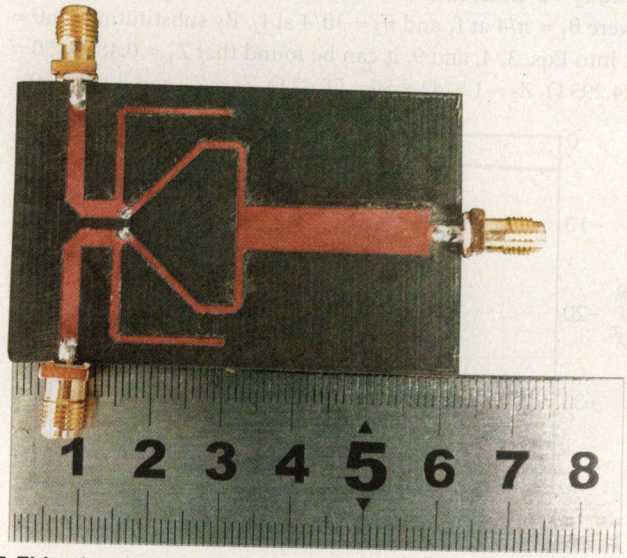
By substituting Eqs. 5 and 6 into Eq. 7 and equating the real and imaginary parts, Eq. 7 can be reduced to Eqs. 8 and 9:

$$R = 2 \quad (8)$$

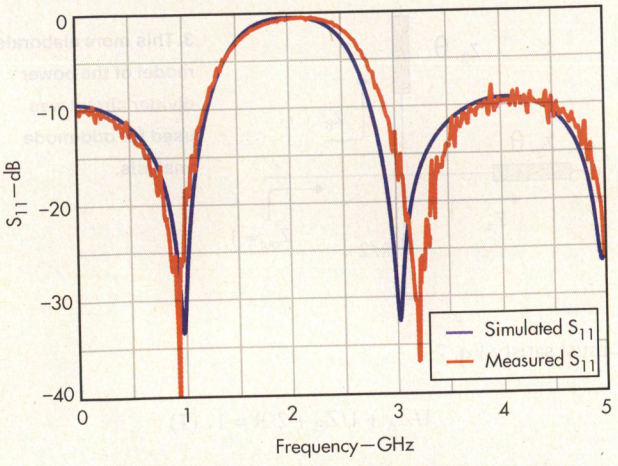
$$\tan^2\theta = Z_3/Z_2 \quad (9)$$

Dual-band analysis was performed on the power-divider design in the following way. Assuming that the center frequencies of the proposed power divider can be found at frequencies f_1 and f_2, and f_2 is greater than f_1, they can be made equivalent by means of the factor k, where $f_2 = kf_1$ ($k > 1$). For the purpose of working in two center frequencies and bands, Eqs. 3, 4, 8, and 9 should be constructed around both center frequencies f_1 and f_2. Eq. 10 should then be satisfied, where:

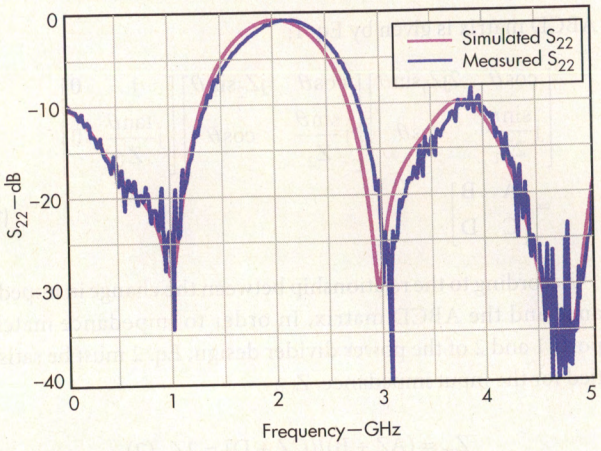
$$\tan^2\theta_{f1} = \tan^2\theta_{f2} \quad (10)$$



5. This photograph shows the size of the fabricated Wilkinson power divider.



6. These curves show the simulated and measured input return loss for the Wilkinson power divider.



7. These curves trace the computer-simulated and measured output return loss for the novel Wilkinson power divider.

From Eq.10, Eqs. 11 and 12 can be obtained for θ:

$$\theta_{f1} = \pi/(1 + k) \quad (11)$$

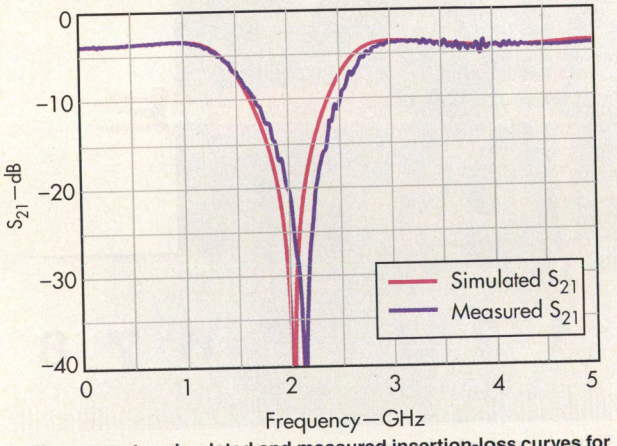
$$\theta_{f2} = k\pi/(1 + k) \quad (12)$$

According to that given operating frequency ratio, k, by solving nonlinear equations Eqs. 3, 4, and 9, the values of impedances Z_1, Z_2, and Z_3 can be obtained. *Figure 4* shows the relationship between impedances Z_1, Z_2, and Z_3 and the frequency ratio k. From *Fig. 4*, it can be seen that the dual-band power divider can be realized within a relatively wide frequency range.

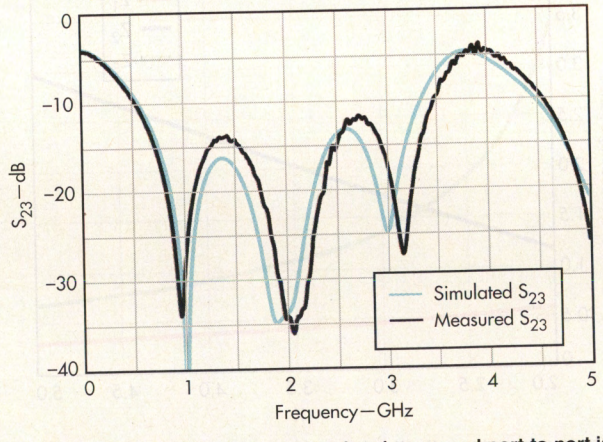
To verify the design method presented for this Wilkinson power divider, an actual power divider was designed and fabricated. Target parameters were set by choosing f_1 = 1 GHz and f_2 = 3 GHz, then k = 3. The electrical lengths of the lines were $\theta_1 = \pi/4$ at f_1 and $\theta_2 = 3\theta/4$ at f_2. By substituting $\tan\theta$ = 1 into Eqs. 3, 4, and 9, it can be found that $Z_1 \approx 0.486 \times 50 = 24.295\ \Omega$, $Z_2 \approx 1.5742 \times 50 = 78.62\ \Omega$, $Z_3 = Z_2$, and $R = 2 \times 50 = 100\ \Omega$. Using a microstrip line calculator, the length and width for each microstrip line needed to achieve these impedances was calculated.

The power divider that was fabricated as a result is shown in *Fig. 5*. In the experiments, RT/5880 circuit-board material from Rogers Corp. (www.rogerscorp.com) was used as the substrate material. This circuit material includes a relative dielectric constant (ε_r) of 2.2 in the z-direction (thickness) with a $\tan\delta$ of 0.0009. The thickness of the substrate dielectric layer was 0.787 mm and the conductor thickness was 0.018 mm.

Scattering parameter measurements were performed using a model N5244A vector network analyzer (VNA) from Agilent Technologies (now Keysight Technologies; www.keysight.com) over the frequency range from 0.02 to 5 GHz. *Figures 6* through 9 show the frequency responses of the divider in which the center frequency of the lower and upper bands were found not to be equal in the simulated and measured situation, respectively. The discrepancies between the simulated and measured results were



8. The computer-simulated and measured insertion-loss curves for the novel Wilkinson power divider can be compared with these plots.



9. These plots compare the simulated and measured port-to-port isolation for the power divider.

mainly caused by the limited accuracy of the simulation processes used.

Moreover, *Figs. 6 and 7* show that the input and output return loss were better than 22 dB at the power divider's two center frequencies. *Figure 8* shows that the insertion loss was less than 0.4 dB across the full frequency range of testing, while *Fig. 9* reveals that the port isolation exceeded 27 dB.

As these different plots of performance indicate, this dual-band Wilkinson power divider provides two clearly defined center frequencies, in terms of both simulated and measured results, at 1 and 3 GHz. Although the computer simulations of input return loss in *Fig. 6* tend to be somewhat idealized in their responses at the two center frequencies, and perfectly positioned on those two center frequencies, the measured results are not far spectrally removed from those two center frequencies and, in fact, register even deeper input return loss than the simulations.

Similarly, the plots of output return loss in *Fig. 7* are even more closely aligned between the simulated and measured results, with both power divider center frequencies revealing 35-dB or better output return loss. Considering how closely the simulate and measured output return loss plots track each other, especially when one considers that measured data at these frequencies can exhibit deviations according to even slight changes in the measurement setup, such as variations in the test connector torque each time the prototype Wilkinson power divider of *Fig. 5* was connected to the VNA for measurements, the output return loss is extremely well behaved for this power divider.

The computer simulated and measured insertion-loss curves of *Fig. 8* show the expected deviations in loss as a function of frequency, but even these are very minimal, with variations in test results also made possible by such things as deviations in connector torque in the test setup. The low insertion loss and high isolation between ports for Wilkinson power divider are sure signs that the design approach should yield similarly satisfactory results at target frequencies other than 1 and 3 GHz and with different circuit-board material with different relative dielectric constant.

In short, this dual-band power divider incorporates stub lines at the ends of the transmission lines for consistent performance over broad frequency ranges. The range of lower and upper band frequency ratios as presented for this power divider design could extend from 2 to 6. An even greater ratio is possible by changing the width of the impedances Z_1, Z_2, and Z_3. The power divider can deliver good performance at any two frequencies.

The design only uses one lumped element and the resistance is 100 Ω at any frequency. Without the use of lumped reactive components, the effects of parasitic circuit elements were minimal, allowing the design to operate at higher frequencies if desired or required. The output port isolation was only related to the length and width of the transmission line Z_2 and the stub line Z_3 relatively. MW

ACKNOWLEDGMENT

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Spectrum and Network Measurements, by Robert A. Witte

SPECTRUM AND NETWORK MEASUREMENTS can be considered the lifeblood of the RF/microwave electronics industry, since these measurements provide tremendous insight into the frequency-domain behavior of devices, components, circuits, and systems at RF, microwave, and millimeter-wave frequencies. The author of this invaluable text, Robert A. Witte, is well known to users of spectrum analyzers and network analyzers from his long career in the test-and-measurement industry and his time spent with the company now known as Keysight Technologies (www.keysight.com), has been a leader in developing high-frequency measurement technologies. The 2nd edition of *Spectrum and Network Measurements* brings Witte's classic text up to date for readers seeking to know about current measurement technologies, such as modular instruments, fast Fourier Transforms (FFTs), and how other instruments fit into a measurement workbench with spectrum and network analyzers.

The clear and concise text includes some early tutorial sections on the meaning of engineering basic parameters, such as the decibel (dB), gain and loss calculations, and how error percentages are denoted in terms of decibels. The book includes chapters on swept spectrum analyzers and their measurements, modulation measurements, distortion measurements, and electromagnetic-compatibility (EMC) measurements.

The opening chapter serves as an introduction to some of the measurement instruments used for spectrum and network analysis showing for example, images of commercial spectrum and network analyzers and simple block diagrams representing the main functions of these instruments. Graphic images are also presented for typical measurement screens, such as a plot of signal amplitude across a frequency bandwidth for a spectrum analyzer, with details on important display parameters, such as resolution bandwidth, video bandwidth, start and stop frequency, and sweep speed. Veteran users of spectrum analyzers certain know the interrelationships of these different spectrum-analyzer measurement parameters. But new users will benefit greatly from reading this text and these introductory chapters to better know the capabilities of modern spectrum and network analyzers. This opening chapter closes with several graphical images and details on some of the newer products in this instrument space, including a combination spectrum and network analyzer that operates on battery power in a compact, portable case, and an example of the latest "rage" in RF/microwave measurement equipment, the PXI mainframe filled with PXI instrument modules. The densely packed capabilities of these PXI instruments allow the equivalent of one or more full racks of equipment to fit into a single 19-in. rack-mount chassis.

An interesting early chapter is devoted to Fourier theory, basically to explain the relationships between signals in the frequency and time domains. This chapter includes the mathematical relationships for Fourier series expansions and uses the Fourier series expansion of a square wave as an example of how to apply these mathematical functions. The text includes details on discrete Fourier transforms and fast Fourier transforms (FFTs), providing brief but effective explanations of how and why they are used in modern measurement setups.

Plots are included to show the effects of different levels of harmonic signals when presented with fundamental test tones, with waveforms that contain third, fifth, seventh, and ninth harmonic signals. The text provides the practical relevance of processing harmonic signals and how the amplitude levels of the harmonics relate to the amplitude levels of the fundamental signals being tested. It even details the case of even-order harmonics, which are rarely mentioned in educational materials, but a legitimate concern. As this chapter concludes, it explains that a spectrum analyzer is capable of measuring the levels of even-order harmonic signals although an oscilloscope cannot.

The chapter on Fourier transforms serves as an excellent lead-in to a chapter on FFT analyzers and how they function in the measurement of high-frequency signals. A simple block diagram is included to illustrate a signal flow diagram from input, through variable attenuator, lowpass filter, sampler, analog-to-digital converter (ADC), and then FFT analysis in a microprocessor and presentation of analyzed signals on the FFT analyzer's display screen. Summation equations are included to describe the role of waveform sampling in capturing and displaying signals in an FFT analyzer, and simple sampling theory is presented to explain the minimum levels of sampling required to accurately represent a captured waveform in digital form.

This chapter includes an explanation of one of the measurement capabilities that has been fairly unique to FFT analyzers, the generation of spectral maps. A graphic is displayed with an oscillator as an example of how spectral maps from an FFT analyzer can be used to analyze harmonic distortion and other oscillator imperfections when evaluating a signal source. The FFT chapter also explains other useful display capabilities, such as histograms and how they are used to show how a signal's amplitude is distributed between its minimum and maximum values. Essentially, a histogram displays the number of samples versus amplitude for an analyzed signal. It can be used for determining the statistical properties of noise and other signal properties, and used to produce analytical tools, such as probability density functions (PDFs) and cumulative density functions (CDFs) that aid in analyzing noise characteristics.

Veteran users of spectrum analyzers will appreciate a chapter devoted to swept

spectrum analyzers and how these measurements have long supported signal analysis in the RF/microwave industry. A swept spectrum analyzer expands the capability of a wave analyzer to measure one frequency at a time and sweeps the analyzer's local oscillator (LO) across a frequency range of interest to measure the amplitude as a function of frequency for that bandwidth. Several simple block diagrams are included for swept spectrum analyzers, with and without microprocessor control, and important details are included on how different parameters, such as sweep speed and LO feedthrough, can affect measurement accuracy. Explanations are provided for the role of the digital intermediate-frequency (IF) section in modern swept spectrum analyzers and the functions of different types of detectors (such as normal, average, and peak detectors) and how they are used for different measurements with a swept spectrum analyzer. The chapter explores the impact of a tracking generator on swept spectrum measurements, and also compares the capabilities of swept spectrum measurements and FFT analyzer measurements. It concludes with an introduction to the real-time spectrum analyzer (RTSA) and what various parameters on an RTSA data sheet, such as probability of intercept (POI), mean in a practical sense.

FURTHER READING

Additional chapters cover modulation measurements, including amplitude modulation (AM) and frequency modulation (FM), quadrature modulation, error vector magnitude (EVM), distortion measurements (including harmonic distortion and intermodulation distortion), noise and noise measurements (including the meaning of equivalent noise bandwidth) and how to make critical noise measurements, such as phase-noise measurements, and pulse measurements, including how to analyze and understand different pulse signal parameters.

This is an ambitious text that covers a great deal of ground related to spectrum and network measurements. It even includes chapters on transmission lines and coaxial connectors, why they are important in modern high-frequency systems, but how such factors as impedance mismatches can impact the accuracy of different spectrum and network measurements.

Obviously, this one book cannot cover all the history and engineering details related to RF/microwave spectrum and network measurements. But it is an excellent starting place. And it includes many helpful hints, such as the use of different components, for improving return loss and other parameters related to enhancing measurement accuracy.

2014, 350 pp., hardcover, ISBN: 978-1-61353-014-6, SciTech Publishing, Edison, NJ, www.scitechpub.com.



TIPS FOR BIASING MMIC AMPLIFIERS

MANY MONOLITHIC-MICROWAVE-INTEGRATED-CIRCUIT (MMIC) amplifiers are shipped without self-biasing networks. Knowing the process for biasing these devices is therefore essential for building quality microwave devices. Biasing techniques differ for each of the three types of MMIC amplifiers: standard, cascode, and hetrojunction bipolar transistor (HBT). In "MMIC Amplifier Biasing Procedure," Hittite Microwave discusses the methods for biasing each variety of MMIC amplifier. The application note also describes the setup required to perform successful biasing.

Two types of HBT MMICs are discussed in the paper: self-biased and self-biased with current control. The self-biased HBT MMIC relies on simple collector voltage activation with a bias resistor. That resistor sets the current for the current mirror, which controls the base voltage. The self-biased HBT MMIC with current control requires an external voltage, known as the power-down control voltage. It controls the current of the current mirror and the base voltage. The paper details the exact sequence needed to ensure proper device operation and prevent critical failures in HBT amplifiers.

For the standard amplifier typology, the bias sequence requires a dual supply, which is positive for the drain-source and negative for the gate-source voltage. The gate voltage is critical in adjusting the drain current for optimal amplifier operation. The process for properly biasing the standard amplifier is detailed in the beginning of the application note, which later tackles power supplies at zero and a damaged-part troubleshooting approach.

It also discusses a cascade-distributed amplifier with a complex biasing structure. This amplifier requires a positive supply for each gate for the unit cell of two FETs in series. Additionally, the amplifier relies on a negative supply for the gate-source voltage. After all the bias interconnections have been set, the first voltage to increase is the gate-supply voltage from the first transistor in the cascode arrangement. Yet it only increases enough to reach pinch-off drain current. If oscillations are observed, the note recommends adding a resistor in series with the gate line or lines.

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CUT THE COST OF WIRELESS MANUFACTURING TEST

FOR MANY YEARS, signaling test was the standard method used in the manufacturing test of wireless devices. This approach involved communicating to the wireless device through functional tests that simulated a wireless network in operation. In recent years, this type of testing has given way to a more efficient and robust test strategy. This new approach also aids in processing devices in more rapid-pace, non-signaling test. Anritsu recently released a white paper titled "Lowering the Cost of Wireless Manufacturing Test," which discusses the use of non-signaling test and test equipment to aid in lowering the cost of manufacturing test for wireless devices.

A non-signaling test setup involves the use of a PC, the device-under-test (DUT), chipset driver/control library, test equipment, and a test executive. The test executive is provided by the chipset company or test vendor or must be produced in-house. Although an in-house solution may offer more customizable control, using a test executive provided by a test vendor may enable a much lower cost and faster setup time for testing. After all, the test executive from the test vendor is more likely to be easily compatible with the test equipment.

The paper discusses the benefits of using a multi-DUT test environment for the rapid and efficient use of testing-equipment resources. For example, the traditional "ping-pong" test approach uses two DUT slots in a testbench and alternates the device being tested while another is set up. Parallel transmit and receive testing works more efficiently for devices that use standards like time division duplexing (TDD), as TDD devices can only transmit or receive at one time. With the advancements of PC-based chassis, which contain multiple test instruments, it is possible to develop a complete multi-DUT test platform in a small footprint using a single-host PC. Because this approach is very flexible, devices that require a dedicated-control PC can be tested efficiently through many of the latest platforms. The key is that Ethernet connectivity has become more available for this type of equipment.

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RC-1SPDT-A18	1	0.25	1.2	85	10	485.00
RC-2SPDT-A18	2	0.25	1.2	85	10	785.00
RC-3SPDT-A18	3	0.25	1.2	85	10	1080.00
RC-4SPDT-A18	4	0.25	1.2	85	10	1280.00
RC-8SPDT-A18	8	0.25	1.2	85	10	2595.00

*The mechanical switches within each model are offered with an optional 10 year extended warranty. Agreement required. See data sheets on our website for terms and conditions. Switches protected by US patents 5,272,458; 6,650,210; 6,414,577; 7,633,361; 7,843,289; and additional patents pending.

†See data sheet for a full list of compatible software.

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Special Feature

JACK BROWNE | Technical Contributor

AWG Delivers 20 GHz with

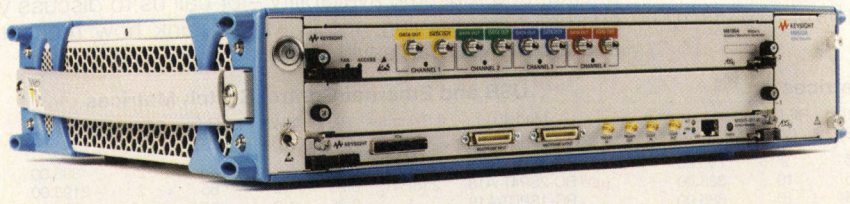
Packed within a compact AXIe module, this arbitrary waveform generator offers 8-b vertical resolution and as many as four synchronized channels to produce complex modulated signals.

Signals required for testing RF/microwave components, devices, and systems grow more complex with time, especially as the demands for wireless services grow and the limits on available bandwidth become apparent. Test signals must include sophisticated forms of modulation based on amplitude, frequency, phase, and even pulses. Fortunately, arbitrary waveform generators (AWGs) provide the means to transform requests for complex test signals into stable, low-noise outputs.

The new M8195A AWG from Keysight Technologies boasts some of the most outstanding performance levels yet to be achieved by these generators, with frequency bandwidth to 20 GHz using an 8-b, 65-GSamples/s architecture packed into a compact, one-slot AXIe module. Each one-slot module contains four digital-to-analog converters (DACs) operating at a maximum rate of 65 GSamples/s, with each M8195A AWG module capable of four differential output signals. This is a test signal source that can match the most advanced produce the most creative forms of modulation and even frequency-hopped signals with minimal jitter. It is a test signal source that can match the most advanced waveforms and the most creative forms of modulation—even frequency-hopped signals—with minimal jitter.

The M8195A AWG *(Fig. 1)* is available with one, two, or four differential channels per one-slot AXIe module; the module can be installed in a two-slot or five-slot AXIe chassis. The 65-GSamples/s sample rate can be maintained on all channels across an instantaneous bandwidth of DC to 20 GHz. The AWG features built-in frequency and phase calibration to produce wideband, multitone signals with flat frequency response to 20 GHz.

The phase-noise and spurious noise levels of these signals are low, but outputs can also be programmed with required amounts of noise and distortion for realistic testing of analog, digital, and optical systems. The M8195A can be programmed and controlled by means of a soft-front-panel application running on the AXIe embedded controller or, alternately, software running on an external personal computer (PC).



1. The M8195A arbitrary waveform generator packs as many as four differential channels with 65 GSamples/s sampling rate and 20-GHz bandwidth into a single AXIe module.

In addition, the M8195A can be controlled by SCPI and IVI-COM remote programming interfaces. The AWG is designed to work with a number of external software programs, including MATLAB from The MathWorks (www.mathworks.com) and LabView from National Instruments (www.ni.com).

This is a competitive signal source for the many challenges of testing modern optical, RF/microwave, and digital systems. It can provide four independent analog output signals, allowing for advanced optical-communications component and system testing, as well as for driving optical systems that require a pair of independent in-phase/quadrature (I/Q) dual-polarized signals. The four output channels can be precisely synchronized in time, with femto-second resolution.

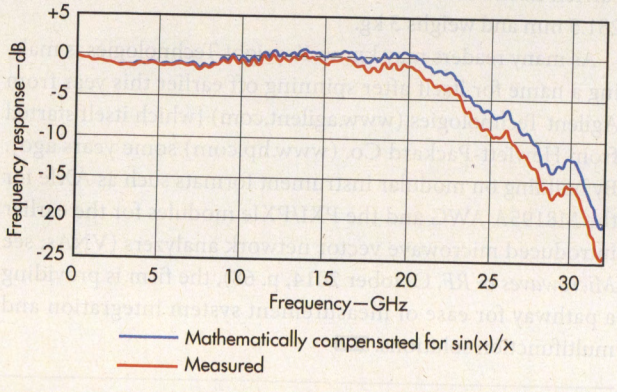
65 GSamples/s

Modern optical communications systems are running with increasingly complex modulation formats as the amount of data to be handled increases, including quadrature-phase-shift-keying (QPSK) modulation, 16-state quadrature amplitude modulation (QAM16), and orthogonal-frequency-division-multiplex (OFDM) modulation. The M8195A can generate all of these modulation formats and more as required for optical communications testing.

In addition, the M8195A is a capable RF/microwave signal generator, offering wideband multitone signals and modulation formats to 20 GHz (*Fig. 2*; the available frequency range depends on the number of channels in use and the amount of available memory per channel). It can generate continuous-wave (CW), pulsed, and frequency-hopped signals with any type of amplitude- or phase-based modulation, such as QAM and PSK.

Because the M8195A attains a high-frequency limit of 20 GHz, it can be used without an external frequency upconverter for many applications, enabling waveform generation directly at the carrier frequency. Frequency switching speed for standard models is better than 505 µs and can be enhanced to 38 ps with frequency-switching option FSW.

In terms of digital signal generation, the M8195A can support testing across a wide range of multiple-level, multiple-channel digital interfaces with almost negligible jitter.



2. The frequency response of the M8195A AWG is relatively flat to 20 GHz.

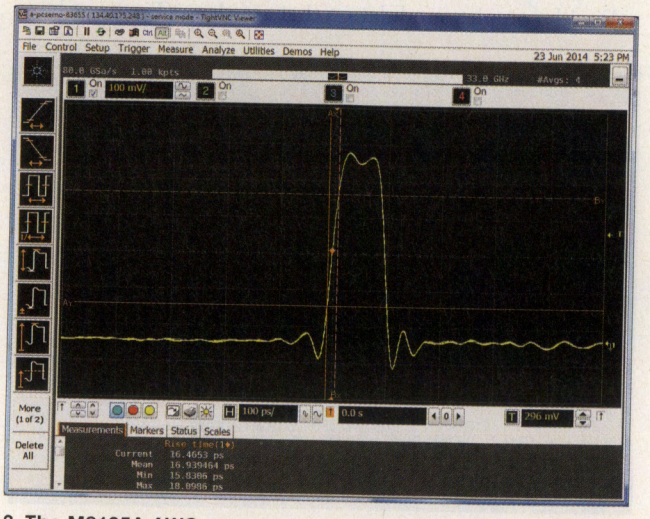
The AWG includes digital predistortion techniques to control the frequency and phase responses of output signals for flexible testing.

For advanced testing, channels can be embedded and de-embedded by knowing the S-parameters of a target circuit or component under test and using these S-parameters to program the M8195A. The AWG offers excellent time-domain behavior for pulsed outputs, with typical 20%/80% rise/fall time of 18 ps *(Fig. 3)*.

The M8195A can load new waveforms on the fly without interrupting the playback of waveforms in progress. Each AXIe module can pack as much as 16 GSamples/s of sample waveform memory (as an option), with 2 GSamples/s as a standard starting point. With its capability for multiple-module synchronization, as many as 16 synchronized channels can be produced from a five-slot AXIe chassis using four M8195A AWG modules.

PLOTTING PERFORMANCE

In terms of performance, the M8195A AWG offers outstanding spectral purity by merit of its 8-b vertical resolu-



3. The M8195A AWG was used to produce this 100-ps pulse with 17-ps rise time.

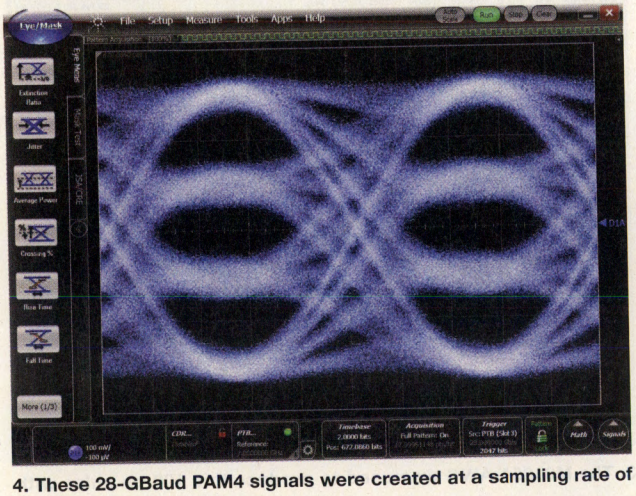
> "The M8195A can generate continuous-wave (CW), pulsed, and frequency-hopped signals with any type of amplitude- or phase-based modulation, such as QAM and PSK."

tion and 65 GSamples/s maximum sampling rate. It achieves an in-band spurious-free dynamic range (SFDR) of typically −80 dBc for outputs from DC to 1 GHz, −53 dBc from 4 to 8 GHz, −50 dBc from 8 to 10 GHz, −46 dBc from 10 to 12 GHz, and −42 dBc or better at frequencies above 12 GHz. Second-harmonic distortion is typically −45 dBc for output signals under 3 GHz, −35 dBc from 3 to 8 GHz, and −30 dBc above 8 GHz. Third-harmonic distortion is typically −35 dBc under 2 GHz and −30 dBc for output frequencies and 2 GHz and higher.

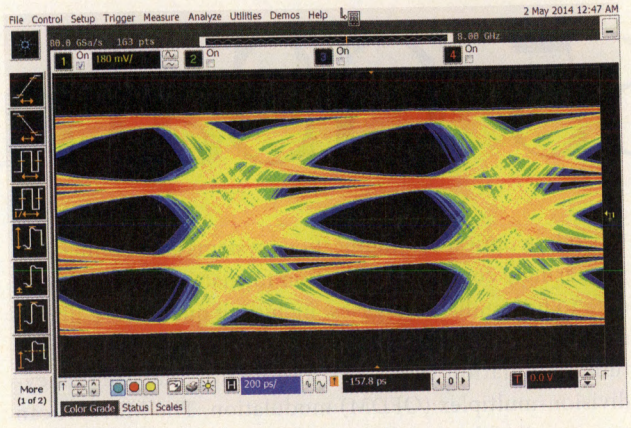
The phase noise for a 1-GHz carrier at 64 GSamples/s is −110 dBc/Hz offset 1 kHz from the carrier, −123 dBc/Hz offset 100 kHz from the carrier, and −140 dBc/Hz offset 10 MHz from the carrier. The time-domain performance is similarly impressive, with skew (peak-to-peak jitter) between any pair of outputs of 0 ± 5 ps at a sampling rate of 32 Gb/s per signal. Residual jitter is less than 200 fs for 32-GBaud signals.

The M8195A also offers outstanding amplitude performance with single-ended outputs to +1 V peak to peak (pp) and differential outputs to +2 V pp across a voltage window of −1.0 to +3.3 VDC. The nominal amplitude resolution is 400 μV and the amplitude accuracy is typically ±2.5% ± 10 mV. The DC offset accuracy is typically ±20 mV.

The AWG maintains flat frequency response across its wide bandwidth with the aid of an integrated 16-tap finite-impulse-response (FIR) filter. The amplitude flatness is typically ±2 dB from DC to 10 GHz, and typically within +2 to −3 dB from 10 to 20 GHz. This is an AWG that can generate signals with complex modulation that are clean *(Fig. 4)* or have



4. These 28-GBaud PAM4 signals were created at a sampling rate of 56 Gb/s.



5. As this screen shows, 28-GBaud PAM4 signals can also be produced with controlled distortion.

controlled amounts of distortion *(Fig. 5)*.

The M8195A offers an adjustable delay range for when compensation must be added for the electrical lengths of components such as coaxial cables. All four channels can be set with as much as 10 ns time delay, adjustable with 50-fs resolution and with accuracy of ±10 ps. These variable delay capabilities also can be used to synchronize the channels of multiple M8195A modules in time.

Each M8195A AWG module sports a reference-clock input port that accommodates signals from 10 MHz to 17 GHz. It can be used as the reference input for all channels in a synchronous system. The internal synthesizer clock of the M8195A has accuracy of ±2 ppm from 54 to 65 GHz with phase noise of better than −115 dBc/Hz offset 10 kHz from a 1-GHz carrier and better than −95 dBc/Hz offset 10 kHz from a 10-GHz carrier. Each M8195A AXIe module measures 322.25 × 30 × 281.5 mm and weighs 3 kg.

As many readers may know, Keysight Technologies is making a name for itself after spinning off earlier this year from Agilent Technologies (www.agilent.com) [which itself started from Hewlett-Packard Co. (www.hp.com) some years ago]. By building on modular instrument formats such as AXIe for the M8195A AWG and the PXI/PXIe modules for the earlier introduced microwave vector network analyzers (VNAs; see *Microwaves & RF*, October 2014, p. 68), the firm is providing a pathway for ease of measurement system integration and multifunction solutions. mw

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	ZVE-3W-183+	5900-18000	35	2	3	1295
NEW	ZHL-4W-422+	500-4200	25	3	4	1570
NEW	ZHL-5W-422+	500-4200	25	3	5	1670
	ZHL-5W-2G+	800-2000	45	5	6	995
	ZHL-5W-1	5-500	44	8	11	1020
	ZHL-10W-2G	800-2000	43	10	13	1295
•	ZHL-16W-43+	1800-4000	45	13	16	1595
•	ZHL-20W-13+	20-1000	50	13	20	1395
•	ZHL-20W-13SW+	20-1000	50	13	20	1445
	LZY-22+	0.1-200	43	16	32	1495
	ZHL-30W-262+	2300-2550	50	20	32	1995
	ZHL-30W-252+	700-2500	50	25	40	2995
	LZY-2+	500-1000	47	32	38	2195
	LZY-1+	20-512	42	37	50	1995
•	ZHL-50W-52+	50-500	50	40	63	1395
•	ZHL-100W-52+	50-500	50	63	79	1995
•	ZHL-100W-GAN+	20-500	42	79	100	2395
	ZHL-100W-13+	800-1000	50	79	100	2195
	ZHL-100W-352+	3000-3500	50	100	100	3595
	ZHL-100W-43+	3500-4000	50	100	100	3595
	LZY-5+	0.4-5	52.5	100	100	1995

Listed performance data typical, see minicircuits.com for more details.

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Cover Story

JACK BROWNE | Technical Contributor
JEAN-JACQUES DELISLE | Technical Engineering Editor

TOP PRODUCTS of 2014

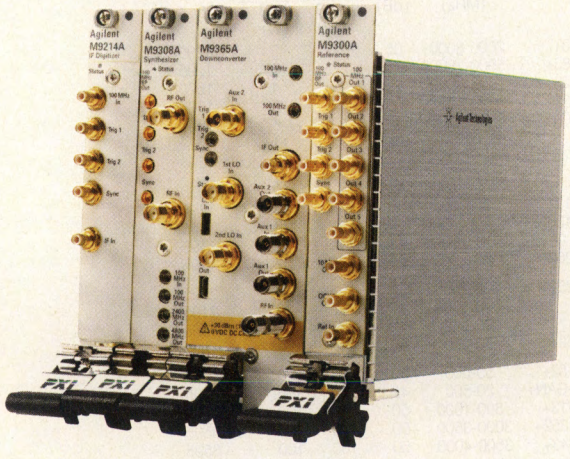
Product advances during the year benefit customers of active and passive components, as well as test equipment, serving applications from RF through millimeter-wave frequencies.

ANY LIST THAT attempts to highlight the "top" products for a particular market during a calendar year will be somewhat arbitrary and a cause for debate, since engineers will show interest in different products depending on their needs. But some types of products that are required fairly regularly, such as building-block components and test equipment, can usually be found on the *Microwaves & RF* Top Products of the Year list for a given year, and the same is true for 2014.

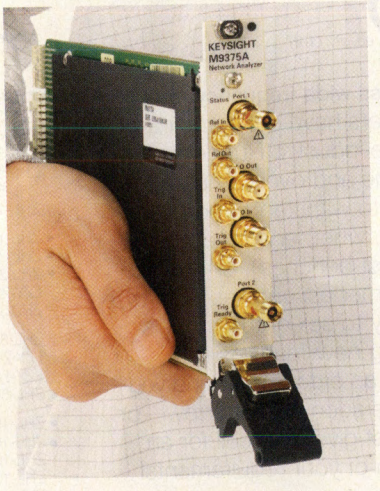
It has been a year of solid RF/microwave engineering and technical advances as the industry continues to find practical ways to generate, send, and receive electromagnetic (EM) energy across wider bandwidths and at higher frequencies.

Building on its introduction of RF/microwave vector signal analyzers (VSAs) in the compact PXI Express (PXIe) module format during 2013, Agilent Technologies (www.agilent.com) extended its PXIe VSA product line in 2014 with the high-performance model M9393A PXIe VSA. Later in the year, the firm would spin off the part of the company devoted to practical measurement solutions as a separate public company called Keysight Technologies (www.keysight.com), and would release yet another line of products that would land on the *Microwaves & RF* Top Products of 2014 list.

Many of the products known to users as Agilent Technologies' test instruments, such as spectrum analyzers and vector network analyzers (VNAs), would become Keysight Technologies' products. For example, the M9393A VSAs introduced earlier in the year as Agilent Technologies' products are now Keysight Technologies' products. The full Top Products of 2014 list is shown in the accompanying *table*.



1. Using the modular PXIe format, the M9393A vector signal analyzer (VSA) can be quickly reconfigured for different frequency ranges and measurement capabilities. *[Photo courtesy of Agilent Technologies (www.agilent.com).]*

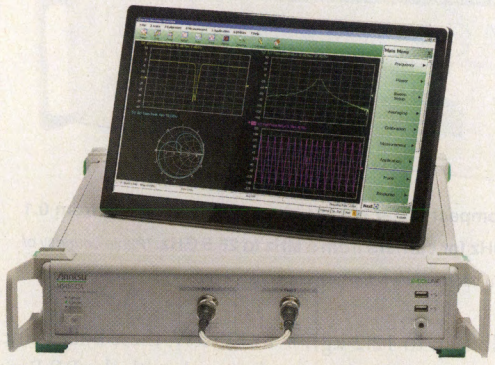


2. The M937XA series of PXI/PXIe vector network analyzer (VNA) modules can serve applications as wide as 300 kHz to 26.5 GHz. *[Photo courtesy of Keysight Technologies (www.keysight.com).]*

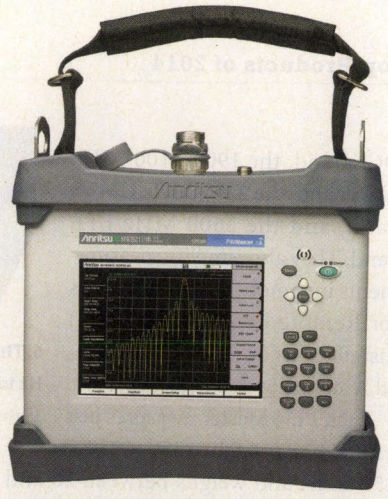
The M9393A PXIe VSA (*Fig. 1*) allows users to assemble test systems with different modules, for a measurement range of 9 kHz to 27 GHz with the most broadband modules—or test bandwidths of 9 kHz to 8.4 GHz, 9 kHz to 14.0 GHz, and 9 kHz to 18.0 GHz with other modules. For a model M9393A VSA, PXIe modules fill five slots in a PXIe chassis and include a frequency downconverter, a frequency reference, a digitizer/intermediate-frequency (IF) processor, and a microwave frequency synthesizer.

When used with the firm's model M9037A PXIe embedded controller, which is essentially a personal computer in PXIe for-

mat, the VSA can achieve its top signal switching speed of 150 µs. The VSAs offer many options in terms of analysis bandwidths and memory storage. The standard analysis bandwidth is 40 MHz, with options for 100 and 160 MHz. The standard instrument memory is 128 Msamples, with options for 512 and 1024 Msamples of memory.



3. The ShockLine VNAs offer economical amplitude and phase measurements to 40 GHz. *[Photo courtesy of Anritsu Co. (www.anritsu.com).]*



4. The portable PIM Master testers simplify in-field testing of distortion in wireless communications systems. *[Photo courtesy of Anritsu Co. (www.anritsu.com).]*

Keysight also appears on the Top 2014 Products list for its M937XA series of PXI/PXIe VNA modules for applications as wide in bandwidth as 300 kHz to 26.5 GHz (*Fig. 2*). Each PXI/PXIe module packs two high-density printed circuit boards (PCBs) and two ports of microwave VNA measurement capability, making it possible to fill a PXI/PXIe chassis with as many as 16 measurement modules and 32 fully synchronized VNA measurement ports.

Examples of the bandwidths covered by these modules include 300 kHz to 4.0 GHz in the model M9370A, 300 kHz to 9.0 GHz in the model M9372A, and 300 kHz to 26.5 GHz in the model M9375A module. For ease of use, the VNA modules work with the same proven user interface and standard commands for programmable instruments (SCPI) programming commands as the benchtop PNA series of RF/microwave VNAs introduced earlier by Agilent (and now supported by Keysight).

A somewhat older name in RF/microwave test-and-measurement technology, Anritsu Co. (www.anritsu.com), landed several products on the Top Products list for 2014, including its ShockLine Series of VNAs for users on tight budgets. Available in two- and four-port versions with frequency coverage as wide as 1 MHz to 40 GHz in two-port systems and as wide as 50 kHz to 8.5 GHz in four-port systems, these VNAs are simple to use and simple in format, with manual operation possible by connection to a touchscreen (*Fig. 3*). They are well suited for performing full S-parameter measurements on passive components, with two-port dynamic range of better than 110 dB and corrected directivity of better than 42 dB.

The low-cost instruments deliver accurate results with fast measurement speeds, operating at sweep speeds as fast as 70 µs/point thanks to patented VNA-on-chip technology. The compact VNAs can fit into 2U-high chassis and can use a connection to a local-area-network (LAN) interface for remote control. Each of these low-cost VNAs is shipped with a standard three-year warranty, even at a U.S. price tag starting at just under $13,000.

The firm also earned Top Products recognition for its MW82119B passive-intermodulation (PIM) Master line of portable testers for measurements from 2 MHz to 3 GHz. These analyzers can run on battery power and still deliver the 40-W test-signal tones needed to produce PIM in the field. These portable units (*Fig. 4*) combine the high-power test-signal generators with the capabilities of the firm's popular Site Master cable and antenna analyzers, enabling contractors and service providers to meet many of their in-field RF/microwave communications system measurement needs without hauling multiple test instruments.

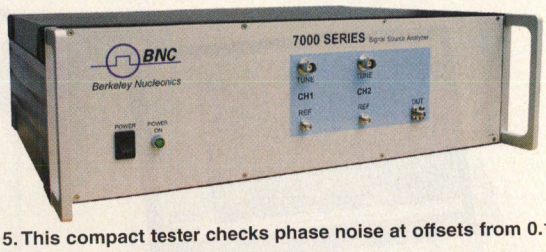
The PIM Master testers can be specified for popular wireless bands, such as the 900 MHz E-GSM band, the 1800-MHz

TOP PRODUCTS OF 2014 (LISTED ALPHABETICALLY)
Agilent Technologies' (now Keysight Technologies') M9393A PXIe VSA (*MW&RF*, April, p. 76)
Anritsu Co.'s economical ShockLine VNAs (*MW&RF*, July, p. 84)
Anritsu Co.'s battery-powered portable PIM Master testers (*MW&RF*, November, p. 97)
Berkeley Nucleonics' model 7300 phase-noise test system (*MW&RF*, March, p. 80)
Freescale Semiconductor's RF Power Tool for amplifier optimization (*MW&RF*, August, p. 91)
Keysight Technologies' M937XA series of PXI/PXIe VNA modules (*MW&RF*, October, p. 86)
Mini-Circuits' model TSS-53LNB+ LNA with integral bypass protection (*MW&RF*, February, p. 74)
Mini-Circuits' model MDB-73H+ SMT double-balanced mixer (*MW&RF*, July, p. 89)
Peregrine Semiconductor's reconfigurable front-end chip (*MW&RF*, March, p. 74)
Synergy Microwave Corp.'s surface-mount DROs (*MW&RF*, May, p. 114)
Tektronix' MDO3000 series of multiple-function, mixed-domain oscilloscopes (*MW&RF*, April, p. 82)
TriQuint Semiconductor's high-power GaN-on-SiC MMIC switch (*MW&RF*, October, p. 100)

DCS band, the 1900/2100-MHz AWS band, the 2100-MHz UMTS band, and the 2600 MHz LTE band. They also simplify measurements in remote locations, such as for Remote Radio Head (RRH) installations or indoor Distributed Antenna Systems (DAS).

Each PIM Master measures just 13.8 × 12.4 × 6.0 in. (350 × 314 × 152 mm) and weighs between 20 and 27 lb (9.0 to 12.2 kg). Each runs on a rechargeable lithium-ion battery with about three hours run time per charge. They feature 8.4-in (213-mm) diagonal color touchscreen displays with 800 × 600 pixel resolution. Screen settings can be adjusted for cloudy conditions and night-time use. To help with the first time in the field, line-sweep tools (LST) software for managing test results and one-day training course are available for learning more about PIM measurements.

Another important high-frequency measurement addressed by a new test product in 2014 was phase noise, and the model 7300 phase-noise test system from Berkeley Nucleonics (www.berkeleynucleonics.com) provides a variety of noise and other measurements on carriers from 5 MHz to 26.5 GHz. The compact, single-box solution *(Fig. 5)* can evaluate phase-noise levels at frequency offsets from 0.1 Hz to 1 MHz.



5. This compact tester checks phase noise at offsets from 0.1 Hz to 1 MHz for carriers from 5 MHz to 26.5 GHz. *[Photo courtesy of Berkeley Nucleonics (www.berkeleynucleonics.com).*

The tester includes an internal frequency reference source but can also be used with external frequency reference sources from 5 MHz to 6 GHz and power levels from +8 to +18 dBm. It can work with internal frequency references from 2 MHz to 26 GHz. The instrument exhibits its residual phase-noise floor of −140 dBc/Hz offset 1 Hz from the carrier, −150 dBc/Hz offset 10 Hz from the carrier, -160 dBc/Hz offset 100 Hz from the carrier, −175 dBc/Hz offset 1 kHz from the carrier, and −180 dBc/Hz offset 10 kHz and more from the carrier.

A firm usually associated more with active devices than measurements, Freescale Semiconductor (www.freescale.com), made a real contribution to developers of RF power amplifiers with their RF Power Tool. The compact instrument *(Fig. 6)*—which works with one of the company's evaluation boards, a power supply, and a dummy load—can make all the key pulsed and CW amplifier measurements from 1 to 2500 MHz (with 1-Hz resolution), including gain, efficiency, bandwidth, VSWR, and bias requirements.

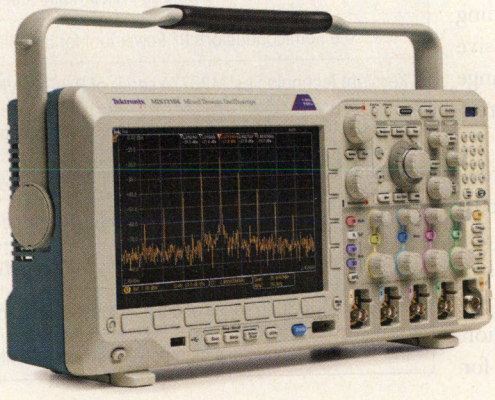
In support of the tester, the company offers evaluation boards for amplifier power levels from 50 W to 125 kW; these boards include sensors and other connectors that work with the RF Power Tool to complete the measurements. The RF Power Tool is connected to a PC and can work with software on the computer, including LabVIEW test software from National Instruments (www.ni.com).

In terms of measurement power for the dollar, few can match Tektronix (www.tektronix.com) during 2014 with the introduction of its MDO3000 series of mixed-domain oscilloscopes (MDOs). Designed almost like a test laboratory with a handle, each MDO3000 instrument contains an oscilloscope, spectrum analyzer, logic analyzer, protocol analyzer, arbitrary function generator, and digital voltmeter *(Fig. 7)*.

The oscilloscopes can be equipped with two or four measurement channels and measurement bandwidths from 100 MHz to 1 GHz while the spectrum analyzers can display test signals as high as 3 GHz. The oscilloscopes include a great variety of trigger functions and, for those in need of test signal flexibility, the optional arbitrary function generator can work at output rates to 250 MSamples/s to generate sine waves to 50 MHz and pulsed signals to 25 MHz. For added flexibility, a built-in waveform editor and integrated software provide 13 predefined waveform functions.



6. The RF Power Tool helps amplifier developers optimize their designs through 2.5 GHz. *[Photo courtesy of Freescale Semiconductor (www.freescale.com).*



7. The MDO3000 series of mixed-domain oscilloscopes combine many useful test tools, including oscilloscopes and spectrum analyzers. *[Photo courtesy of Tektronix (www.tektronix.com).*

CREATING COMPONENTS

For users of low-noise amplifiers (LNAs), one of the more clever product innovations introduced during 2014 was in the form of the surface-mount model TSS-53LNB+ LNA from Mini-Circuits (www.minicircuits.com). While LNAs are essen-

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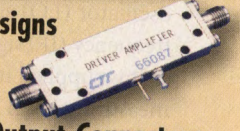



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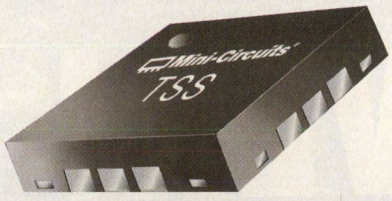
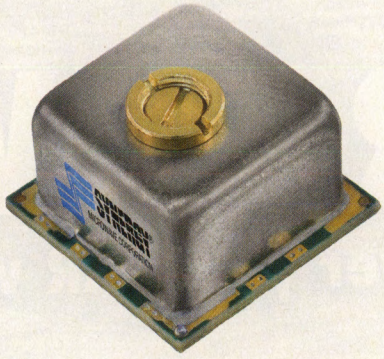
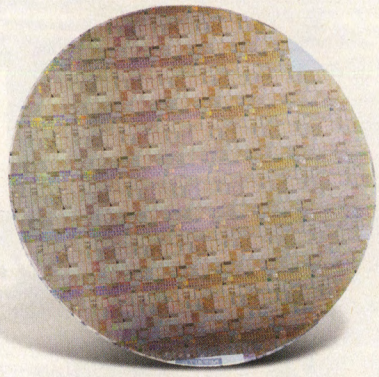


8. This miniature LNA operates to 5 GHz and includes bypass protection. *[Photo courtesy of Mini-Circuits (www.minicircuits.com).]*



9. This compact housing is the surface-mount home for low-noise DROs capable of fundamental-frequency outputs to 6 GHz and higher. *[Photo courtesy of Synergy Microwave Corp. (www.synergymwave.com).]*



10. A reconfigurable front-end design helps push silicon CMOS technology to higher frequencies. *[Photo courtesy of Peregrine Semiconductor Corp. (www.psemi.com).]*

tial for boosting low-level signals for such applications as receivers, they are also often victims of unexpected large-level signals. This 0.5-to-5.0-GHz LNA features a switchable bypass circuit with minimal signal loss, which can be selected in the presence of large signals to protect the LNA's sensitive input port.

Based on 0.25-µm enhancement-mode GaAs pseudomorphic high-electron-mobility-transistor (pHEMT) semiconductor technology, the LNA provides reliable operation while drawing about 80 mA typical current from a +5-VDC supply. It provides about 21.5-dB small-signal gain and 1.3-dB typical noise figure at 2 GHz while maintaining gain flat within ±0.7 dB from 0.7 to 2.1 GHz without need for impedance-matching components. This RoHS-compliant LNA *(Fig. 8)* is supplied in a surface-mount 12-lead 3 × mm MCLP package with on-die electrostatic-discharge (ESD) protection.

During 2014, Mini-Circuits also continued to advance one of the technologies with which it is most associated, the microwave mixer. The company introduced the model MDB-73H+ double-balanced mixer, its first monolithic-microwave-integrated-circuit (MMIC) mixer. Based on indium-gallium-arsenide (InGaP) semiconductor technology, the mixer works with RF and local-oscillator (LO) signals from 2.2 to 7.0 GHz and intermediate-frequency (IF) signals from DC to 1.6 GHz.

It is supplied in a surface-mount-technology (SMT) package measuring just 4 × 4 × 1 mm. With its outstanding performance and operating temperature range of –40 to +85°C, the mixer is a candidate for both commercial and military broadband systems.

Synergy Microwave Corp. (www.synergymwave.com) has rarely failed to make some form of advancement in RF/microwave source technology during a given year and 2014 was no different for the firm. Synergy's development of low-noise, surface-mount-device (SMD) dielectric resonator oscillators (DROs) aimed for outstanding performance in compact housings.

These sources *(Fig. 9)* employ planar resonators based on metamaterial Mobius strips to achieve extremely low phase-noise levels at fundamental frequencies to 6 GHz and beyond. They also boast the low thermal frequency drift so essential for many applications across various markets.

Synergy's SMD DROs are supplied in housings measuring just 0.75 × 0.75 in. and are also available on printed metamaterial Mobius strips in 0.5 × 0.5 in. and 0.3 × 0.3 n. square packages for low-cost, low-noise applications. The compact signal sources can readily be designed and supplied for applications to 18 GHz and beyond.

TriQuint Semiconductor (www.triquint.com) provided an example of its diverse foundry services and capabilities by means of its latest process, gallium nitride on silicon carbide (GaN-on-SiC) technology. When used with high-frequency semiconductor materials such as GaN, the SiC substrate material can effectively channel heat away from active devices. The combination was used in the design of the model TGS2355-SM MMIC single-pole, double-throw (SPDT) switch with CW power-handling capability of 40 W from 500 MHz to 6 GHz and pulsed power-handling capability of 100 W across the same frequency range.

The switch is fabricated on the firm's 0.25-µm production semiconductor process and is supplied in an air-cavity, lead-free, RoHS-compliant quad-flat-no-leads (QFN) package measuring just 5 × 5 × 1.42 mm. At the time of introducing the GaN switch, the firm also revealed that it would merge with RF Micro Devices, Inc. to form the firm known as Qorvo, Inc.

At the same time, Peregrine Semiconductor Corp. (www.psemi.com) showed that a firm without a semiconductor foundry could still be creative and productive, with the introduction of its reconfigurable RF front-end (RFFE) integrated circuit (IC) and its 130-ns device features. The chip *(Fig. 10)*, which provides gallium-arsenide-like performance in silicon, builds upon the company's UltraCMOS Global 1 silicon CMOS technology to densely pack active and passive circuit functions, such as amplifiers and switches, into tight semiconductor spaces.

The technology, which enables improved yields and operating efficiencies compared to GaAs, has already been applied to the fabrication or high-frequency, broadband circuits covering 0.1 to 6.0 GHz and more. mw

50 MHz to 26.5 GHz

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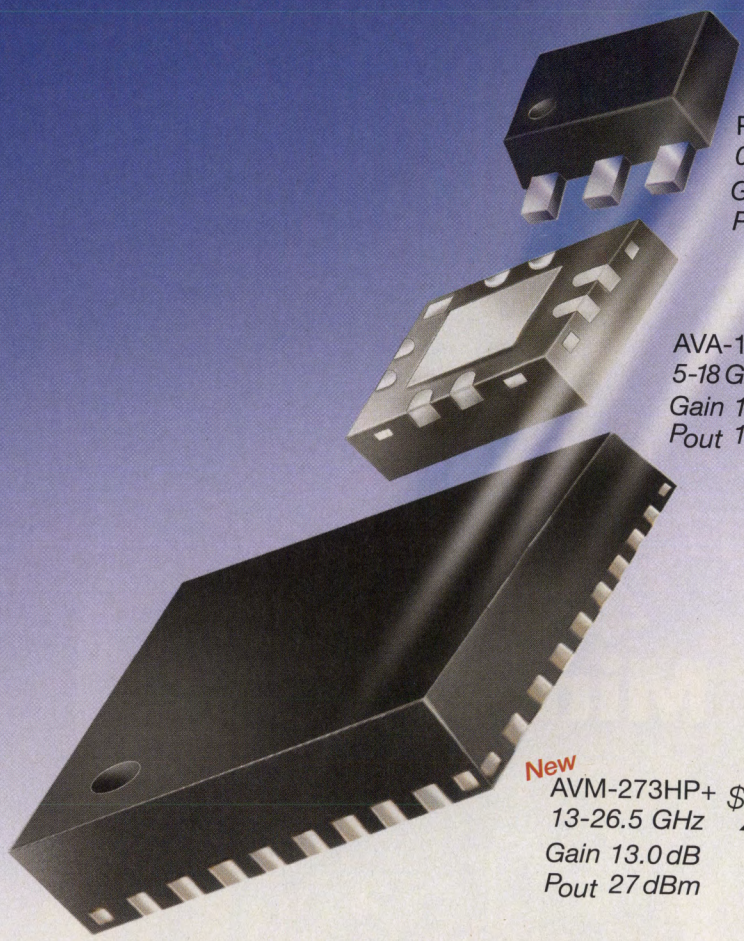

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Mini-Circuits' New AVM-273HP+ wideband, 13 dB gain, unconditionally stable microwave amplifier supports applications from 13 to 26.5 GHz with 0.5W power handling! Gain flatness of ±1.0 dB and 58 dB isolation make this tiny unit an outstanding buffer amplifier in P2P radios, military EW and radar, DBS, VSAT, and more! Its integrated application circuit provides reverse voltage protection, voltage sequencing, and current stabilization, all in one tiny package!

The AVA-183A+ delivers 14 dB Gain with excellent gain flatness (±1.0 dB) from 5 to 18 GHz, 38 dB isolation, and 19 dBm power handling. It is unconditionally stable and an ideal LO driver amplifier. Internal DC blocks, bias tee, and microwave coupling capacitor simplify external circuits, minimizing your design time.

The PHA-1+ + uses E-PHEMT technology to offer ultra-high dynamic range, low noise, and excellent IP3 performance, making it ideal for LTE and TD-SCDMA. Good input and output return loss across almost 7 octaves extend its use to CATV, wireless LANs, and base station infrastructure.

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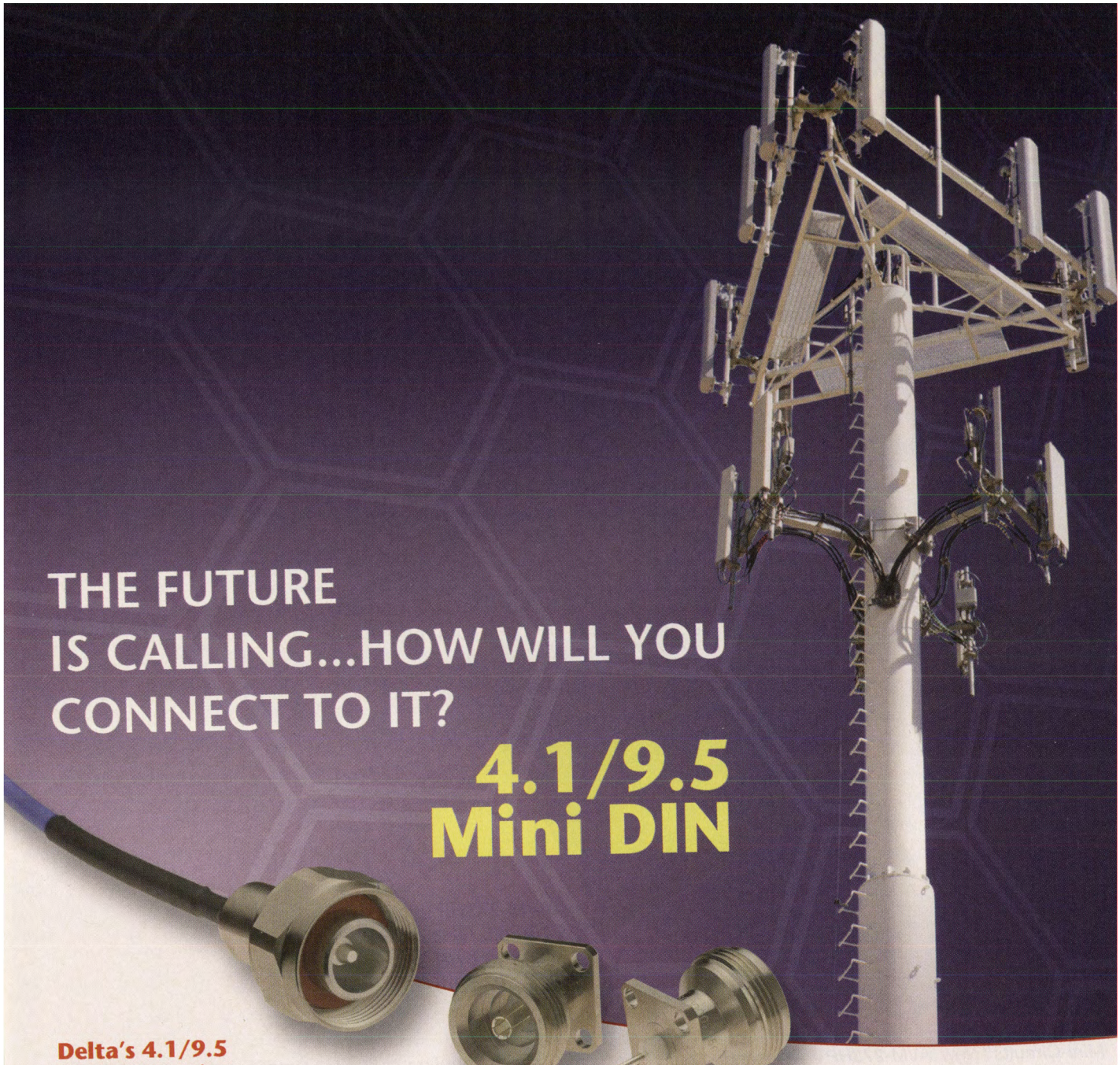

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Real-Time Scope Spans 100 GHz

This high-performance, high-speed oscilloscope boasts performance levels to real-time bandwidth of 100 GHz at 240-GSamples/s sampling rate, with the flexibility to add test capability as needed.

DIGITAL OSCILLOSCOPES HAVE long been measurement workhorses for high-frequency electronics, especially as they have moved higher in frequency. The new LabMaster 10-100Zi real-time oscilloscope from Teledyne LeCroy (www.teledynelecroy.com) is both versatile and powerful, with a real-time analog bandwidth of 100 GHz that is being strobed at sample rates to 240 GSamples/s. The oscilloscope can be run in a number of different multiple-channel modes—as well as in conjunction with computer-aided-test (CAT) and computer-aided-engineering (CAE) software programs—for flexible measurements of the most complex analog and digital signals.



1. The LabMaster 10-100Zi oscilloscope is a modular measurement system with bandwidth to 100 GHz per channel and sampling rate to 240 GSamples/s. *(Screenshots courtesy of Bell Labs, Alcatel-Lucent)*

The LabMaster 10-100Zi real-time oscilloscope *(Fig. 1)* builds upon the modular architecture of the company's lower-frequency LabMaster 10 Zi oscilloscopes, using a two-housing format to deliver a great deal of flexibility. A basic LabMaster 10-100Zi includes a LabMaster MCM-Zi Master Control Module and a LabMaster 10-100Zi Acquisition Module. Modules of each kind can be added as required to form an almost unlimited number of different measurement configurations. Examples include a real-time oscilloscope with multiple 100-GHz channels and a test system with various combinations of 36-, 65-, and 100-GHz measurement channels.

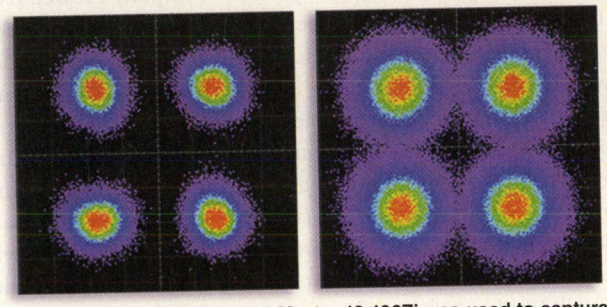
The modular system can be equipped with as many as 20 acquisition modules under the command of a single MCM-Zi control module. It can then form an oscilloscope with as many as 80 measurement channels, each operating at a real-time bandwidth of 36 GHz and sampling rate of 80 GSamples/s; as many as 20 measurement channels, each at a real-time bandwidth of 100 GHz and sampling rate of 240 GSamples/s; and anything in between. At the maximum bandwidth, the oscilloscope exhibits less than 130 fs RMS jitter between channels. Communication between each of the acquisition modules and the CPU is performed by means of high-speed PCIe connections.

Each control module includes a display, operating controls, central processing unit (CPU), and distributed system clock—a highly stable 10-GHz source that maintains precise synchronization among all of the oscilloscope channels. Digital PCIe data connections from the control module to the LabMaster MCM-Zi Acquisition Modules provide transfer of data and control signals. Although each oscilloscope system is capable of acquiring enormous amounts of data, the design employs the firm's pro-

prietary X-Stream II streaming technology to process and manage large amounts of data without compromising data precision and integrity.

The 240-GSamples/s digitizing speed of the LabMaster 10-100Zi real-time oscilloscope means that test samples are acquired at time intervals as brief as 4.17 ps, allowing capture of extremely short events like laser and radar pulses. As communications systems grow in complexity, the timing and synchronization between channels become more finely tuned, requiring more sophisticated measurement capabilities as provided by the LabMaster 10-100Zi. Patented innovations such as digital bandwidth interleaving (DBI) and ChannelSync multiple-module synchronization ensure that separate modules function as a single unit.

Through their use, measurement modules can be added with confidence as channels are needed for various tasks—such as analyzing the complex signals found in digitally modulated signals with in-phase (I) and quadrature (Q) signal components—to a top limit of 80 synchronized measurement channels (each with a real-time bandwidth of 36 GHz). As noted earlier, the usable real-time measurement bandwidth is a function of the number of channels, with as many as 20 real-time measurement channels available at the 100-GHz bandwidth.



2. A two-channel, 100-GHz LabMaster 10-100Zi was used to capture 80 and 160 GBaud QPSK signals.

The ChannelSync architecture features impressive synchronization, with less than 130 fs channel-to-channel jitter. The oscilloscope draws this tight timing from a microwave reference source, a 10-GHz ChannelSync distributed clock that accounts for the precise synchronization among the as many as 80 measurement channels.

The clock accuracy is better than 0.1 ppm + aging rate of 0.1 ppm/yr from the last calibration. The sample clock jitter, for acquired time durations to 3.2 ms, is only 50 fs RMS. For twice the acquired time, to 6.4 ms, the sample clock jitter increases to 130 fs RMS.

The LabMaster 10-100Zi offers wide ranges of time/division scales, depending on measured bandwidth. For measurement bandwidths greater than 36 GHz, time/division scales from 10 ps/div to 640 μs/div. For measurement bandwidths less than or equal to 36 GHz, the time/division scale extends from 10 ps/div to as long as 256 s/div (for truly long events or widely spaced events).

Jitter between channels, measured at the maximum bandwidth, is less than 130 fs RMS. The trigger and interpolator jitter is typically less than 0.1 ps RMS when software assisted, and typically 2 ps as a result of the hardware alone.

The LabMaster 10-100Zi features flexible triggering capabilities, with the firm's SMART Triggers that enable triggering on any input source if it can be state or edge qualified. The instrument system can trigger on signal event widths, with widths selectable from 200 ps to 20 s, or on intermittent faults, triggering on intervals between 1 ns and 20 s. The oscilloscope system supports maximum trigger rate of 1 million waveforms/s for as many as four channels.

A high-speed oscilloscope system such as the LabMaster 10-100Zi can capture and display the growingly complex signals of optical and digital communications systems, such as these 80 and 160 GBaud quadrature-phase-shift-keying (QPSK) signals *(Fig. 2)*. These measurements were made at Alcatel-Lucent's famed Bell Labs (www.alcatel-lucent.com), using with a two-channel LabMaster 10-100Zi system with 100-GHz real-time bandwidth.

Because custom analysis often must be performed on captured high-speed test signals, the LabMaster 10-100Zi teams with a number of different software tools. The high-speed oscilloscope is equipped with the XDEV software package from XDEV Software Corp. (www.xdev-software.com) that allows users to run custom MATLAB code from The MathWorks (www.mathworks.com) for general-purpose signal analysis.

When more specific analysis is needed (for example, examining high-speed optical signals), software developed for optical modulation analysis can just as easily be used with the LabMaster 10-100Zi instruments. For instance, the SDAIII-CompleteLinQ software from Coherent Solutions (www.coherent-solutions.com) can be used with the LabMaster 10-100Zi for eye, jitter, and noise analysis on as many as four signal lanes simultaneously.

Physically, the LabMaster 10-100Zi master control module measures 10.9 × 18.2 × 15.6 in. (277 × 462 × 396 mm) and weighs 47 lbs (21.4 kg). The master acquisition module measures 8.0 × 18.2 × 26.0 in. (202 × 462 × 660 mm) and weighs 58 lbs (24 kg).

The control module also includes a port for an external 10-MHz timebase reference and provides access to a 10-MHz reference signal (at 50 Ω) for use with other test instruments. P&A: $900,000 and up; stock. MW

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2-WAY								
CSBK260S	20 - 600	0.28 / 0.4	0.05 / 0.4	0.8 / 3.0	25 / 20	1.15:1	50	377
DSK-729S	800 - 2200	0.5 / 0.8	0.05 / 0.4	1 / 2	25 / 20	1.3:1	10	215
DSK-H3N	800 - 2400	0.5 / 0.8	0.25 / 0.5	1 / 4	23 / 18	1.5:1	30	220
P2D100800	1000 - 8000	0.6 / 1.1	0.05 / 0.2	1 / 2	28 / 22	1.2:1	2	329
DSK100800	1000 - 8000	0.6 / 1.1	0.05 / 0.2	1 / 2	28 / 22	1.2:1	20	330
DHK-H1N	1700 - 2200	0.3 / 0.4	0.1 / 0.3	1 / 3	20 / 18	1.3:1	100	220
P2D180900L	1800 - 9000	0.4 / 0.8	0.05 / 0.2	1 / 2	27 / 23	1.2:1	2	331
DSK180900	1800 - 9000	0.4 / 0.8	0.05 / 0.2	1 / 2	27 / 23	1.2:1	20	330
3-WAY								
S3D1723	1700 - 2300	0.2 / 0.35	0.3 / 0.6	2 / 3	22 / 16	1.3:1	5	316
4-WAY								
CSDK3100S	30 - 1000	0.7 / 1.1	0.05 / 0.2	0.3 / 2.0	28 / 20	1.15:1	5	169S

▫ With matched operating conditions

HYBRIDS RoHS

Model #	Frequency (MHz)	Insertion Loss (dB) [Typ./Max.] ◊	Amplitude Unbalance (dB) [Typ./Max.]	Phase Unbalance (Deg.) [Typ./Max.]	Isolation (dB) [Typ./Min.]	VSWR (Typ.)	Input Power (Watts) [Max.]	Package
90°								
DQS-30-90	30 - 90	0.3 / 0.6	0.8 / 1.2	1 / 3	23 / 18	1.35:1	25	102SLF
DQS-3-11-10	30 - 110	0.5 / 0.8	0.6 / 0.9	1 / 3	30 / 20	1.30:1	10	102SLF
DQS-30-450	30 - 450	1.2 / 1.7	1 / 1.5	4 / 6	23 / 18	1.40:1	5	102SLF
DQS-118-174	118 - 174	0.3 / 0.6	0.4 / 1	1 / 3	23 / 18	1.35:1	25	102SLF
DQK80300	800 - 3000	0.2 / 0.4	0.5 / 0.8	2 / 5	20 / 18	1.30:1	40	113LF
MSQ80300	800 - 3000	0.2 / 0.4	0.5 / 0.8	2 / 5	20 / 18	1.30:1	40	325
DQK100800	1000 - 8000	0.8 / 1.6	1 / 1.6	1 / 4	22 / 20	1.20:1	40	326
MSQ100800	1000 - 8000	0.8 / 1.6	1 / 1.6	1 / 4	22 / 20	1.20:1	40	346
MSQ-8012	800 - 1200	0.2 / 0.3	0.2 / 0.4	2 / 3	22 / 18	1.20:1	50	226
180° (4-PORTS)								
DJS-345	30 - 450	0.75 / 1.2	0.3 / 0.8	2.5 / 4	23 / 18	1.25:1	5	301LF-1

◊ In excess of theoretical coupling loss of 3.0 dB

COUPLERS RoHS

Model #	Frequency (MHz)	Coupling (dB) [Nom]	Coupling Flatness (dB)	Mainline Loss (dB) [Typ./Max.]	Directivity (dB) [Typ./Min.]	Input Power (Watts) [Max.] ▫	Package
KFK-10-1200	10 - 1200	40 ±1.0	±1.5	0.4 / 0.5	22 / 14	150	376
KDS-30-30	30 - 512	27.5 ±0.8	±0.75	0.2 / 0.28	23 / 15	50	255 *
KBS-10-225	225 - 400	10.5 ±1.0	±0.5	0.6 / 0.7	25 / 18	50	255 *
KDS-20-225	225 - 400	20 ±1.0	±0.5	0.2 / 0.4	25 / 18	50	255 *
KBK-10-225N	225 - 400	10.5 ±1.0	±0.5	0.6 / 0.7	25 / 18	50	110N *
KDK-20-225N	225 - 400	20 ±1.0	±0.5	0.2 / 0.4	25 / 18	50	110N *
KEK-704H	850 - 960	30 ±0.75	±0.25	0.08 / 0.2	38 / 30	500	207
SCS100800-10	1000 - 8000	10.5 ±1.5	±2.0	1.2 / 1.8	8 / 5	25	361
KBK100800-10	1000 - 8000	10.5 ±1.5	±2.0	1.2 / 1.8	8 / 5	25	322
SCS100800-16	1000 - 7800	16.8 ±1.5	±2.8	0.7 / 1.0	14 / 5	25	321
KDK100800-16	1000 - 7800	16.8 ±1.5	±2.8	0.7 / 1.0	14 / 5	25	322
SCS100800-20	1000 - 7800	20.5 ±2.0	±2.0	0.45 / 0.75	12 / 5	25	321
KDK100800-20	1000 - 7800	20.5 ±2.0	±2.0	0.45 / 0.75	14 / 5	25	322
KEK-1317	13000 - 17000	30 ±1.0	±0.5	0.4 / 0.6	30 / 15	30	387

* Add suffix - LF to the part number for RoHS compliant version.
▫ With matched operating conditions

Unless noted, products are RoHS compliant.

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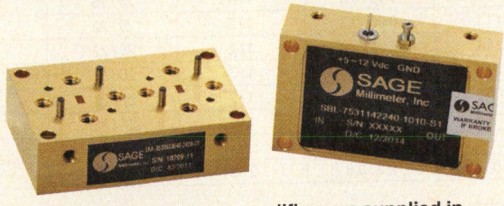
Amplifiers Curb Noise of Millimeter Waves

This pair of LNAs supports a number of communications and radar applications from 60 to 90 GHz and 75 to 110 GHz.



1. These millimeter-wave amplifiers are supplied in compact housings with waveguide connectors.

MILLIMETER-WAVE LINKS ARE expanding as a means of making broadband, high-data-rate connections from one point to another, but those links depend on reliable, high-performance components like low-noise amplifiers (LNAs). Fortunately, Sage Millimeter has built a solid reputation for its high-frequency amplifiers.

The latest pair of LNAs from the supplier combine for total frequency coverage from 60 to 110 GHz with generous gain and low noise figures. Both amplifiers are designed for efficient current consumption from bias supplies of +5 to +12 VDC, and both are supplied in compact housings (*Fig. 1*).

Generous bandwidth made available at higher frequencies—such as the 60 to 90 GHz of the model SBL-6039033050-1212-S1 amplifier and the 75 to 110 GHz of the model SBL-7531142240-1010-S1 LNA—support effective wireless communications links over short, line-of-sight distances.

As more electronic systems move towards wireless communications and control, the flexibility and efficiency of millimeter-wave links shows them to be practical, high-data-rate communications solutions capable of enduring even hostile weather and environmental conditions.

The lower-frequency amplifier, model SBL-6039033050-1212-S1, covers a frequency range (60 to 90 GHz) that makes it a candidate not only for communications receivers and links, but also for radar systems and test equipment. It provides the relatively flat gain expected from an amplifier for test applications, with gain flatness of ±4 dB across the full 30-GHz band, from a minimum level of 26.2 dB to a maximum of 34.1 dB gain (typical gain of 30.0 dB) from 60 to 90 GHz (*Fig. 2*).

The model SBL-6039033050-1212-S1 LNA operates on a single positive bias supply of 30 mA at +5 to +12 VDC. It includes WR-12 waveguide with UG-387/U flange for the input and output signal ports and a solder pin for the bias connection.

For those seeking higher-frequency coverage, the model SBL-7531142240-1010-S1 LNA operates across full W-band frequencies, with impressive gain flatness of typically ±2 dB across the full band from 75 to 110 GHz (*Fig. 3*).

The typical noise figure is 4 dB from 75 to 110 GHz, while the saturated output power reaches typically +2 dBm. This LNA is outfitted with WR-10 waveguide with UG-387/U-M flange for the input and output ports.

Both amplifiers are constructed with right-angle input and output waveguide connectors in standard units, although other mechanical versions of the amplifiers are available. mw

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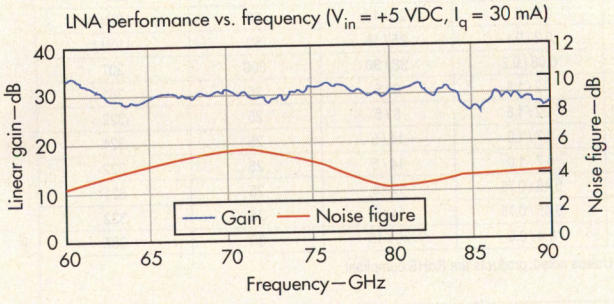
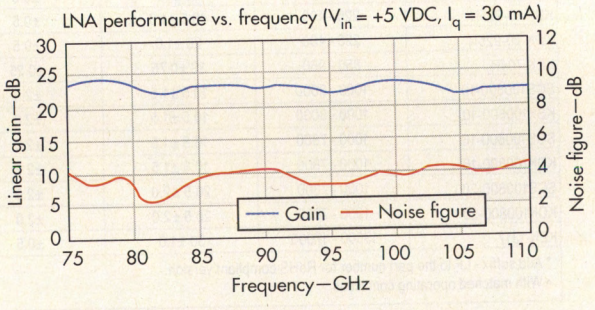


2. The model SBL-6039033050-1212-S1 LNA provides 30-dB typical gain from 60 to 90 GHz.



3. The model SBL-7531142240-1010-S1 LNA offers 22-dB typical gain from 75 to 110 GHz.

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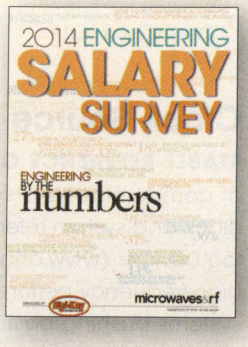

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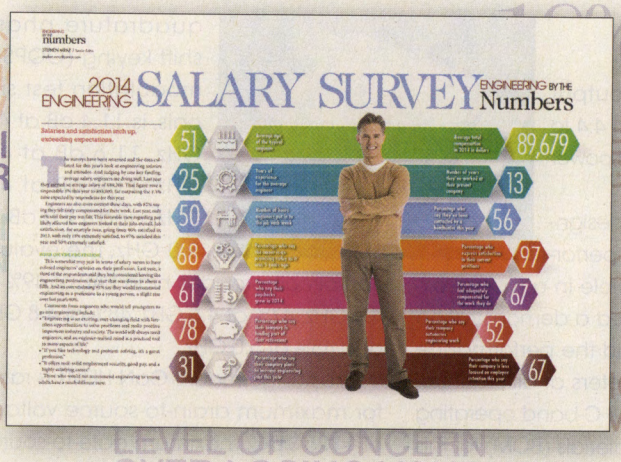

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Clock Source Kills Jitter to 1.25 GHz

STABLE REFERENCE CLOCKS are critical for many forms of communications systems, and the AD9528 JESD204B clock and system reference (SYSREF) generator from Analog Devices (www.analog.com) provides the high-speed/high-frequency performance needed for communications systems pushing past data rates of 1.25 GSamples/s. The JESD204B standard is being used increasingly as the interface between digital signal processors (DSPs) and high-speed data converters in gigabit-speed communications systems.



The AD9528 clock integrated circuit (IC) includes the different functions needed to support high-speed system communications, including clock distribution and an on-board voltage-controlled oscillator (VCO). The VCO, which tunes from 3.6 to 4.0 GHz, supports both single-ended and differential receiver connections.

The AD9528 IC contains a two-stage phase-locked loop (PLL) with integrated JED204B SYSREF generator for synchronization of multiple devices. The first PLL stage furnishes input signal reference conditioning by reducing the amount of jitter present on a system clock. The second PLL stage also helps to minimize jitter by providing high-frequency clock signals with low broadband noise. The AD9528 produces two output signals (PLL1 and PLL2) with maximum frequency to 1.25 GHz, and 12 additional low-noise output signals as high as 1 GHz.

Each of the 14 output channels contains a divider that includes coarse digital phase-adjustment capability and analog phase-delay control for manipulation of timing across all 14 output channels.

By using the on-board timer circuitry, the delay control of the PLL1 output port can be controlled in steps as fine as 17 ps; the delay control of the PLL2 output port can be adjusted in steps as fine as 20 ps. The clock output phase noise depends on the mode of operation and frequency.

The device is designed for typical supply of +3.3 VDC, 1635 mW typical power consumption in single-loop-mode operation, and 1675 mW typical power consumption in dual-loop-mode operation, with operating temperature range of –40 to +85°C.

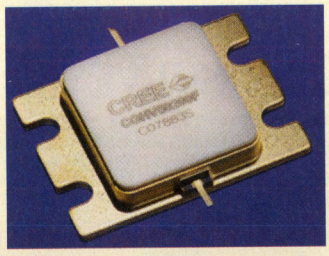
ANALOG DEVICES, INC., Three Technology Way, Norwood, MA 02062; (800) 262-5643, *www.analog.com*

GaN HEMT Packs C-Band Power

C-BAND COMMUNICATIONS SYSTEMS both on the ground and in space rely on generous power to maintain their links. At one time, that signal power was from electron tubes, but now gallium nitride (GaN) high-electron-mobility-transistor (HEMT) active devices such as the model CGHV50200F from Cree (www.cree.com) can deliver tube-like power in packages that are a fraction of the size of electron tubes.

The model CGHV50200F HEMT may fit within a compact ceramic/metal flange package, but it provides 200 W typical continuous-wave (CW) output power from 4.4 to 5.0 GHz. It supports beyond-line-of-sight (BLOS) communications, satellite-communications (satcom), and troposcatter communications applications with high efficiency.

The flange-mount high-power GaN HEMT is internally matched to 50 Ω for ease of installation in many high-frequency circuits and subsystems. It is rated for typical saturated output power of 180 W from 4.4 to 5.0 GHz, with typical power gain of 11.5 dB.



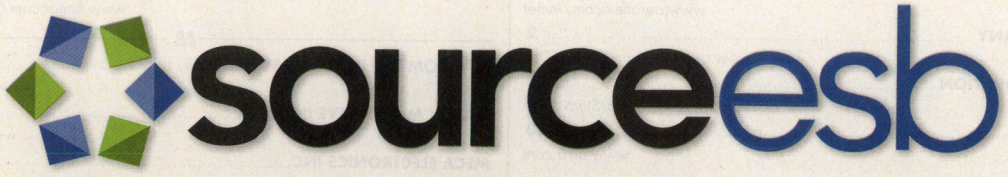
To give amplifier designers an idea of the performance levels possible in a typical circuit, Cree developed a demonstration amplifier and tested the performance for different parameters at different frequencies across the C-band operating range. The bill of materials (BOM) for this demonstration circuit is detailed on the data sheet for the GaN HEMT.

In this circuit, the model CGHV50200F GaN HEMT reaches CW output-power levels of 198 W at 4.4 GHz, 213 W at 4.6 GHz, 233 W at 4.8 GHz, and 218 W at 5.0 GHz. In the same circuit, it achieves small-signal gain of 14.9 dB at 4.4 GHz, 14.9 dB at 4.6 GHz, 14.9 dB at 4.8 GHz, and 15.1 dB at 5.0 GHz.

The power gain in the demonstration circuit, measured with orthogonal quadrature phase shift keying (OQPSK) modulation test signals, is 11.4 dB at 4.4 GHz, 11.6 dB at 4.6 GHz, 11.0 dB at 4.8 GHz, and 11.8 dB at 5.0 GHz. The evaluation circuit provides high PAE of 49% at 4.4 GHz, 47% at 4.6 GHz, 48% at 4.8 GHz, and 48% at 5.0 GHz.

The flange-mount GaN HEMT is rated for maximum drain-to-source voltage of +125 VDC, maximum gate-to-source voltages of +10 and +2 VDC, maximum operating junction temperature of +225°C, maximum forward gate current of 41.6 mA, and maximum drain current of 17 A. The transistor has received an electrostatic discharge (ESD) classification of 1A for the human body model (HBM) and class 2 for the charge device model (CDM).

CREE, INC., 4600 Silicon Dr., Durham, NC 27703; (919) 313-5300, FAX: (919) 869-2733, *www.cree.com/rf*



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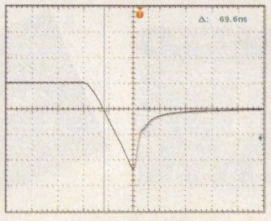

Model AVR-CD1-B output waveform 2 A / DIV 40 ns / DIV



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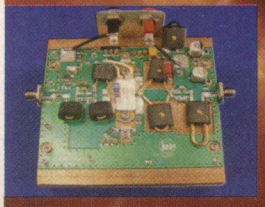

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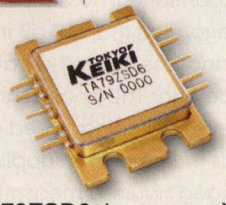

The **TA79ZSD6** is a new X-Band 100W Solid State GaN Power Amplifier. It uses a compact and lightweight 17.4mm x 24mm x 4.1mm package. It has greater than 36% Drain efficiency and 11dB of gain. It requires a +30V supply voltage. All bias, input and output matching circuitry are included.

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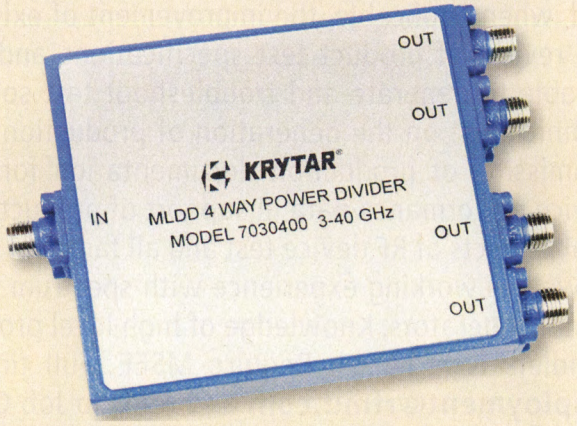
A FOUR-WAY power divider provides the broadband frequency range suitable for test and communications applications, operating from 3 to 40 GHz with low loss. Model 7030400 exhibits less than 3-dB insertion loss from 3 to 32 GHz and less than 3.6-dB insertion loss from 32 to 40 GHz. The isolation is better than 13 dB across the frequency range with ±0.5-dB amplitude tracking. The maximum input VSWR is 1.90:1 from 3 to 32 GHz and 2.50:1 from 32 to 40 GHz. The maximum output VSWR is 1.70:1 from 3 to 32 GHz and 2.0:1 from 32 to 40 GHz. The phase tracking is ±11 deg. from 3 to 32 GHz and ±15 deg. from 32 to 40 GHz. The four-way power divider measures 2.64 × 2.30 × 0.40 in. with 2.4-mm female connectors, although a version is also available with 2.92-mm female K connectors.



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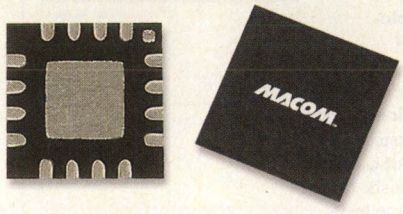
3-GHz PIN Switch Controls 100 W

DESIGNED FOR commercial and military applications requiring robust power-handling capability, model MASW-011030 is a single-pole, three-throw (SP3T) PIN diode switch for applications from 30 MHz to 3 GHz. Designed in a common-anode configuration, the switch can handle 100 W continuous-wave (CW) incident power at a base-plate temperature of +85°C. The compact component achieves

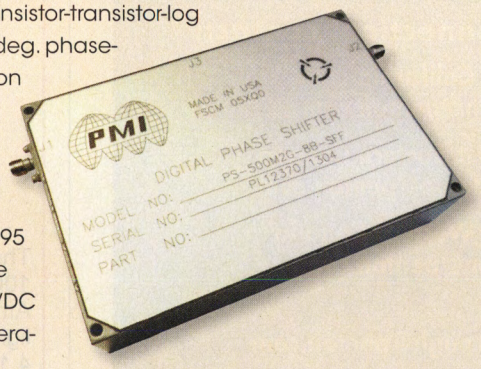


40-dB isolation with only 0.35-dB insertion loss at 2 GHz and more than 15 dB input return loss. It is 1B human-body-model (HBM) electrostatic-discharge (ESD) rated. The switch is supplied in a 7-mm HWFN 16-lead plastic package and is suitable for both high-power CW and pulsed signal operation, including in land-mobile radios and other military communications applications.

MACOM TECHNOLOGY SOLUTIONS HOLDINGS, INC., 100 Chelmsford St., Lowell, MA 01851; (800) 366-2266, (978) 656-2500, FAX: (978) 656-2804; *www.macom.com*

Phase Shifter Controls 0.5 to 2.0 GHz

MODEL PS-500M2G-8B-SFF is a digitally controlled phase shifter with frequency range of 500 to 2000 MHz. It provides 360-deg. phase-shift adjustment over that frequency range via 8-b transistor-transistor-log (TTL) control with less than ±10-deg. phase-shift error, less than 10-dB insertion loss, and less than ±1-dB amplitude error. The phase shifter achieves switching speed of 500 ns or less and is supplied in a rugged housing measuring 4.95 × 3.38 × 1.00 in. with SMA female connectors. It operates on ±15 VDC and is rated for operating temperatures from –55 to +85°C.



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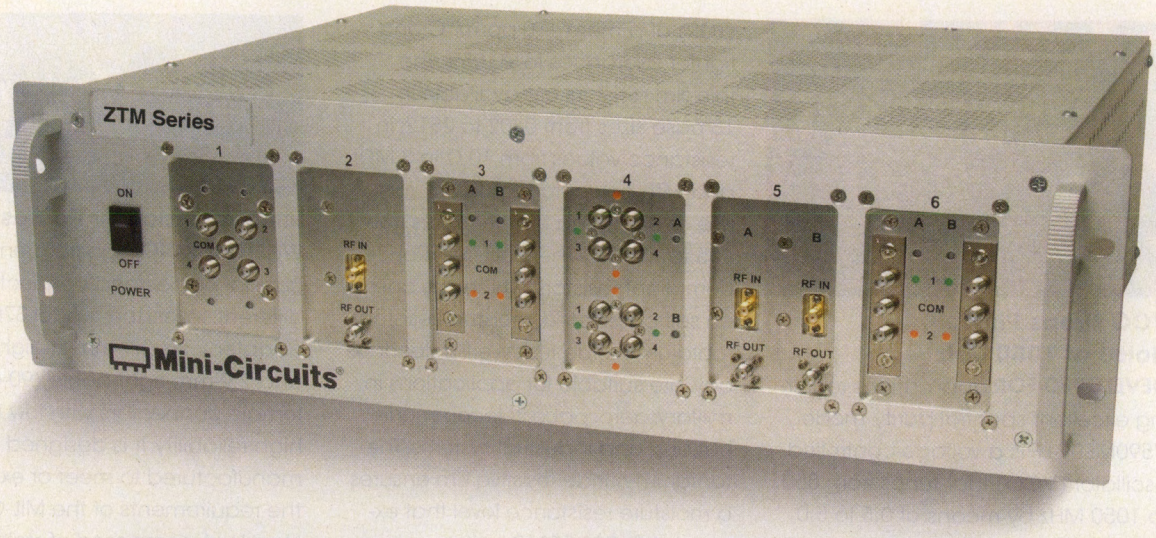
SAW Filter Screens GPS

MODEL AM1575S361 is a surface-acoustic-wave (SAW) filter designed for use in Global Positioning Systems (GPS) at 1575 MHz. The filter offers a 20-MHz bandwidth at that center frequency (passband of 1565 to 1585 MHz) with insertion loss of just 0.9 dB and better than 14-dB return loss. The rejection of unwanted signals and interference is 10 dB at 1525 MHz and 22 dB at 1625 MHz. The filter, which is supplied in a surface-mount-technology (SMT) package format, measures 3.0 × 3.0 × 1.3 mm. It can handle +10 dBm input power.

ANATECH ELECTRONICS, INC., 70 Outwater Lane, Garfield, NJ 07026; (973) 772-4242, FAX: (973) 772-4646, e-mail: sales@anatechelectronics.com, *www.anatechelectronics.com*



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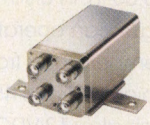

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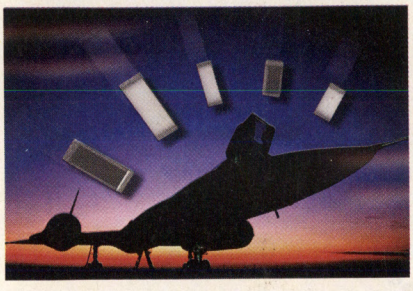
DEVELOPED FOR applications requiring excellent spectral purity, model V590ME09-LF is a voltage-controlled oscillator (VCO) that tunes from 850 to 1050 MHz by means of 0.5 to 5.0 V. It delivers +5.5-dBm output power into a 50-Ω load with average tuning gain of 63 MHz/V/ The phase noise is controlled to a low –100 dBc/Hz offset 10 kHz from the carrier with second-harmonic suppression of –13 dBc, all the while drawing just 15 mA current from a +5-VDC supply. The RoHS-compliant VCO is designed for operating temperatures of –20 to +70°C and is supplied in the firm's standard MINI-14S package measuring 0.5 × 0.5 × 0.22 in. The oscillator is available in tape-and-reel packaging for high-volume production.

Z-COMMUNICATIONS, INC., 14118 Stowe Dr., Ste. B, Poway, CA 92064; (858) 621-2700, FAX: (858) 486-1927, e-mail: applications@zcomm.com, *www.zcomm.com*

Chip Resistors Lower TCRs

THE PTN series of precision surface-mount thin-film chip resistors is now available with a lower absolute temperature coefficient of resistance (TCR) of a mere ±10 ppm/°C and tighter tolerances to ±0.05%. The thin-film resistors are available in 13 case sizes from 0402 to 2512; in resistance values from 10 Ω to 3 MΩ; and with power ratings to 2 W for use at operating temperatures from –55 to +155°C. The lead-free, RoHS-compliant resistors, which are based on self-passivated, moisture-resistant tantalum nitride resistive film technology, are suitable for applications in military, aerospace, telecommunications, and industrial systems. The tantalum-nitride resistive film ensures a moisture resistance level that exceeds MIL-PRF-55342 limits.



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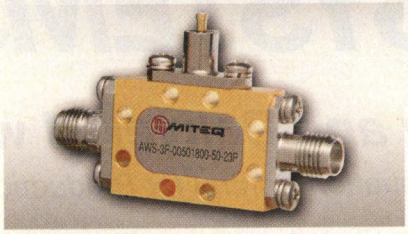

Coaxial Amplifier Gains 0.5 to 18.0 GHz

MODEL AWS-3F-00501800-50-23P is a hermetic, single-bias coaxial amplifier with more than 25-dB small-signal gain from 0.5 to 18.0 GHz. It achieves gain flatness of ±1.5 dB or better across that wide frequency range, with minimum output power of +23 dBm at 1-dB compression. The maximum noise figure is 5 dB. The RoHS-compliant amplifier, which is supplied with field-replaceable SMA connectors, measures just 1.00 × 0.66 in. in a slim-profile Kovar housing. It can operate with base plate temperatures from –40 to +75°C and typically draws 450 mA current from a single +12-VDC supply.

MITEQ, INC., 100 Davids Dr., Hauppauge, NY 11788; (631) 436-7400, FAX: (631) 436-7430, *www.miteq.com*



In-Series Adapter Passes 40 GHz

WHEN LINKING high-frequency coaxial cables and connectors of different sexes, the model 370SF-40 adapter joins 2.9-mm straight male and 2.9-mm right-angle (90-deg.) female connectors with low loss and high reliability. It is designed and manufactured to meet or exceed the requirements of the MIL-C-39012 standard, composed of stainless-steel body and coupling nut per QQ-S-764 Type 303 stainless steel. The contact is formed of beryllium copper as per QQ-C-530, while the insulators are composed of Teflon per MIL-P-19468A. The adapter achieves low VSWR not exceeding 1.30:1 at frequencies through 40 GHz.

UNITED MICROWAVE PRODUCTS, INC., 22129 South Vermont Ave., Torrance, CA 90502; (310) 320-1244, FAX: (310) 320-9729, e-mail: umpinc@pacbell.net, *www.unitedmicrowave.com*

USB Sensor Reads Power to 6 GHz

THE MODEL 7022 statistical RF Universal-Serial-Bus (USB) power sensor measures power levels from 250 mW to 500 W of digitally modulated signals at frequencies from 350 MHz to 6 GHz. Well suited for in-field measurements, the compact unit measures 5.8 × 4.8 × 1.3 in. and weighs just 1.5 lb. In addition to checking power, it can perform a number of other measurements, including VSWR and return loss. It works with the firm's Windows-based Virtual Power Meter software running on a personal computer (PC).

BIRD TECHNOLOGIES, 30303 Aurora Rd., Solon, OH 44139; (440) 248-1200, (866) 695-4569, FAX: (440) 248-5426, *www.birdrf.com*


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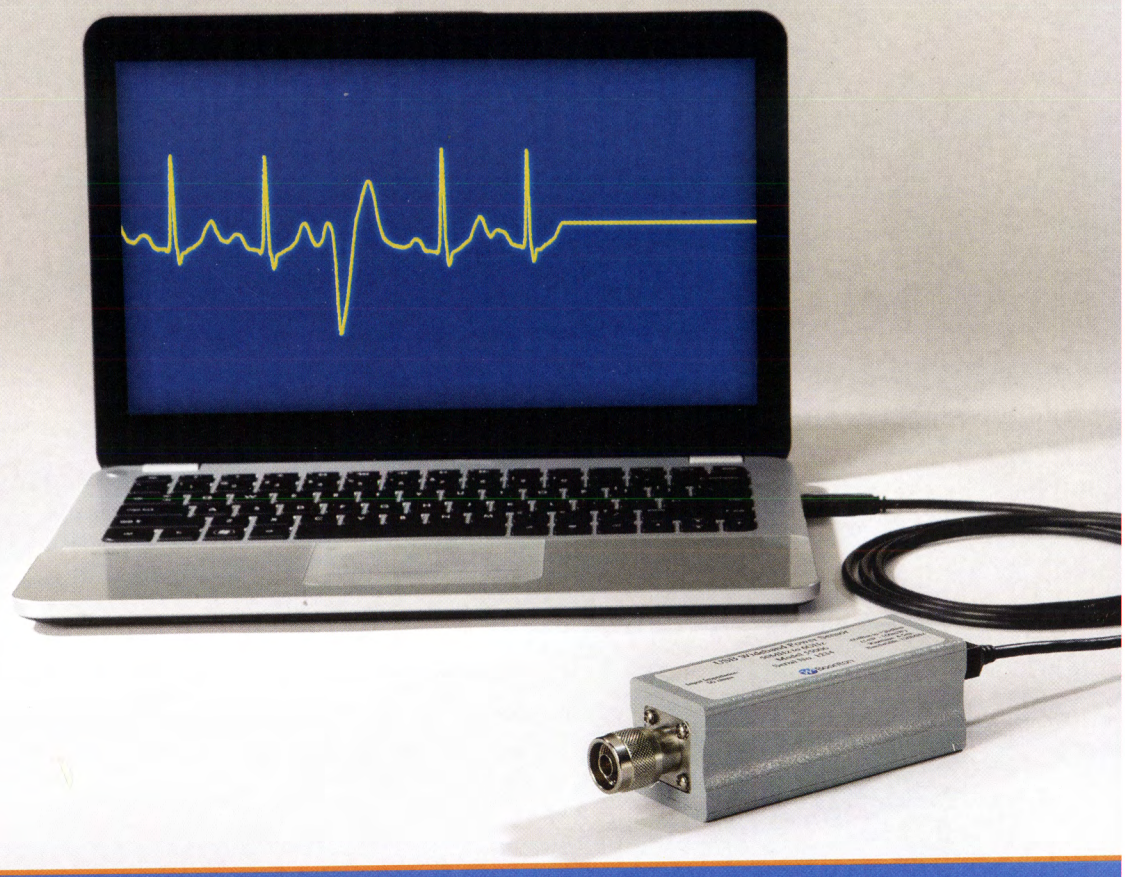

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