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P2T-10M6G-45-R-5V-SFF-HIP20W


P2T-10M8G-45-R-50W-AL


P2T-0R5G8G-45-R-SFF-50W-SM


P2T-500M10G-60-R-515-SFF-10WCW


P2T-1G1R1G-25-R-SFF-100W-SM

PMI Model No.	Frequency Range (GHz)	Insertion Loss (dB Max)	Isolation (dB)	Switching Speed (Typ)	Operating Input Power	DC Voltage & Current (Max)	Size Configuration Connector
P2T-10M6G-45-R-5V-SFF-HIP20W https://www.pmi-rf.com/product-details/p2t-10m6g-45-r-5v-sff-hip20w	10 MHz - 6	2.8	25 Typ	100 ns Max	25 W CW Max	+5 VDC @ 30 mA	1.20” x1.0” x 0.5” SP2T Reflective SMA Female
P2T-10M8G-45-R-50W-AL https://www.pmi-rf.com/product-details/p2t-10m8g-45-r-50w-al	10 MHz - 8	2.6	36 Min	100 ns	50 W CW	+5 VDC @ 100 mA	1.20” x1.0” x 0.5” SP2T Reflective SMA Female
P2T-0R5G8G-45-R-SFF-50W-SM https://www.pmi-rf.com/product-details/p2t-0r5g8g-45-r-sff-50w-sm	0.5 - 8	3.0	30 Min	100 ns	50 W CW	+5 VDC @ 32 mA	1.20” x1.0” x 0.5” SP2T Reflective SMA Female
P2T-500M10G-60-R-515-SFF-10WCW https://www.pmi-rf.com/product-details/p2t-500m10g-60-r-515-sff-10wcw	0.5 - 10	2.5	60 Min	100 ns	10 W CW Max	+ 5 VDC @ 38 mA -15 VDC @ 43 mA	1.20” x 1.00” x 0.5” SP2T Reflective SMA Female
P2T-1G1R1G-25-R-SFF-100W-SM https://www.pmi-rf.com/product-details/p2t-1g1r1g-25-r-sff-100w-sm	1 - 1.1	0.8	25 Min	250 ns	100 W CW, 5 kW Peak	+50 VDC @ 10 mA +5 VDC @ 128 mA	3.25” x 2.75” x 0.7” SP2T Reflective TNC Female
P2T-1G18G-10-R-528-SFF-HIP10W https://www.pmi-rf.com/product-details/p2t-1g18g-10-r-528-sff-hip10w	1 - 18	4.0	10 Min	100 ns	10 W CW	+5 VDC @ 2.19 mA, -28 VDC @ 2.5 mA	1.2” x 1.0” x 0.5” SP2T Reflective SMA Female
P2T-1R2G1R4G-25-R-SFF-250W https://www.pmi-rf.com/product-details/p2t-1r2g1r4g-25-r-sff-250w	1.2 -1.4	0.8	25 Min	250 ns	250 W Peak	+30 V @ 50 mA, +5 V @ 200 mA	4.22” x 2.98” x 0.7” SP2T Reflective SMA Female
P2T-6G18G-40-R-570-TFF-1D6KW https://www.pmi-rf.com/product-details/p2t-6g18g-40-r-570-tff-1d6kw	6 - 18	2.2	40	1.1 µs	1.6 kW Peak - Tested to 125.89 W CW	+ 5 VDC @ 300 mA -70 VDC @ 60 mA	2.0” x 2.0” x 0.75” SP2T Reflective TNC Female
PDT-8G12G-40-515-SFF https://www.pmi-rf.com/product-details/pdt-8g12g-40-515-sff	8 - 12	2.0	40 Min	200 ns Max	40 dBm CW, 57 dBm, 1 µs, 1% DC	+5 VDC @ 100 mA -15 VDC @ 60 mA	1.25” x 1.25” x 0.3” SP2T Reflective SMA Female
P2T-8G18G-50-R-SFFF https://www.pmi-rf.com/product-details/p2t-8g18g-50-r-sfff	8 - 18	3.0	50 Min	200 ns	40 W Max, 300 W Peak	+5 VDC @ 150 mA, -28 VDC @ 80 mA	1.55” x 2.3” x 0.5” SP2T Reflective SMA Female


P2T-1G18G-10-R-528-SFF-HIP10W


P2T-1R2G1R4G-25-R-SFF-250W


P2T-6G18G-40-R-570-TFF-1D6KW


PDT-8G12G-40-515-SFF


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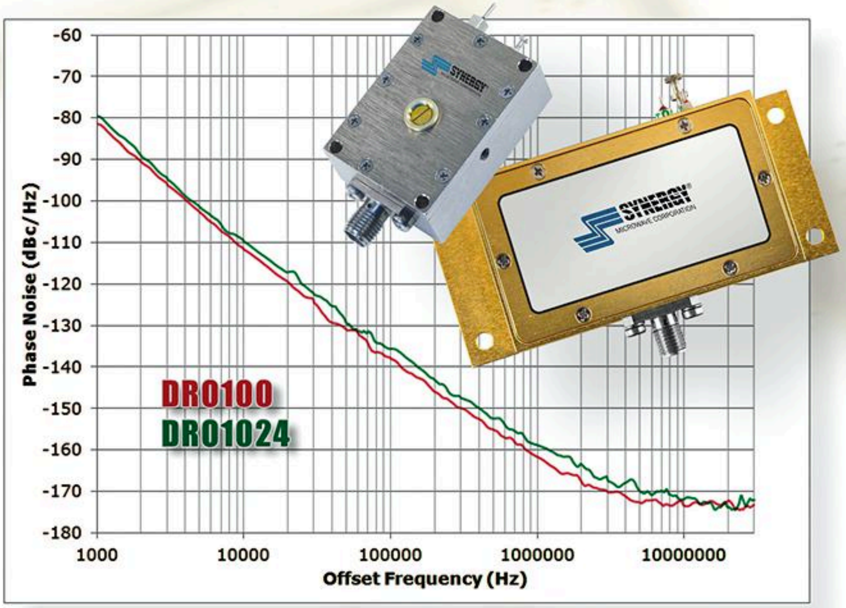

Model	Frequency (GHz)	Tuning Voltage (VDC)	DC Bias (VDC)	Typical Phase Noise @ 10 kHz (dBc/Hz)
Surface Mount Models				
SDRO800-8	8.000	1 - 10	+8.0 @ 25 mA	**-114**
SDRO900-8	9.000	1 - 10	+8.0 @ 25 mA	**-114**
SDRO1000-8	10.000	1 - 15	+8.0 @ 25 mA	**-107**
SDRO1024-8	10.240	1 - 15	+8.0 @ 25 mA	**-105**
SDRO1118-7	11.180	1 - 12	+5.5 - +7.5 @ 25 mA	**-104**
SDRO1121-7	11.217	1 - 12	+5.5 - +7.5 @ 25 mA	**-104**
SDRO1130-7	11.303	1 - 12	+5.5 - +7.5 @ 25 mA	**-104**
SDRO1134-7	11.340	1 - 12	+5.5 - +7.5 @ 25 mA	**-104**
SDRO1250-8	12.500	1 - 15	+8.0 @ 25 mA	**-105**
Connectorized Models				
DRO80	8.000	1 - 15	+7.0 - +10 @ 70 mA	**-114**
DRO8R95	8.950	1 - 10	+7.0 - +10 @ 38 mA	**-109**
DRO100	10.000	1 - 15	+7.0 - +10 @ 70 mA	**-111**
DRO1024	10.240	1 - 15	+7.0 - +10 @ 70 mA	**-109**
DRO1024H	10.240	1 - 15	+7.0 - +10 @ 70 mA	**-115**
KDRO145-15-411M	14.500	*	+7.5 @ 60 mA	**-100**

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Editorial

CHRIS DeMARTINO | Editor
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How Well Do You Know James Clerk Maxwell?

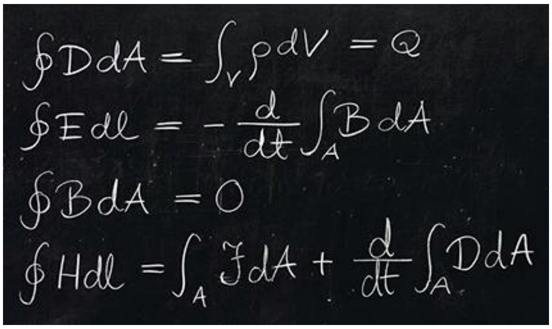
There's no questioning the importance of Maxwell's equations. But do most engineers still remember them?

What are four of the most influential equations in all of science? One could answer that question by saying "Maxwell's equations," which most working RF/microwave engineers should have encountered in an undergraduate course on the road to becoming an engineer. But how many still remember Maxwell's equations in detail? If someone asked you to sit down and write them, would you be able to do it? I will leave you to go get pencil and paper.

While Maxwell's equations form the fundamental principles of electromagnetic (EM) theory, do today's engineers even need to know them? Of course, every person has different responsibilities. But even for those engineers who routinely use EM simulation tools, are Maxwell's equations necessary?

To come up with an answer, let's look at a company that's been in the business of EM simulation for a long time: Sonnet Software *(www.sonnetsoftware.com)*. With a history that dates to 1983, Sonnet is no stranger to EM software. On top of that, the company's founder, James Rautio, has written about James Clerk Maxwell himself (this literature can be found on Sonnet's website).

So, do Sonnet users need to know Maxwell's equations? Brian Rautio, VP of operations at Sonnet, responded, "With over 30 years of development effort, ease of use has been a huge priority for Sonnet, and so knowing Maxwell's equations isn't really a prerequisite for use. The software discretizes the circuit geometry, casts it into a form of Maxwell's equations, and solves the problem with relative fluidity. Indeed, in some cases, the user doesn't even need to enter so much as the permittivity of the material.



$$\oint D\,dA = \int_V \rho\,dV = Q$$

$$\oint E\,dl = -\frac{d}{dt}\int_A B\,dA$$

$$\oint B\,dA = 0$$

$$\oint H\,dl = \int_A J\,dA + \frac{d}{dt}\int_A D\,dA$$

(Continued on page 6)

(Continued from page 5)

"That said," Rautio added, "there are some concepts that are evasive to software automation—for example, accounting for ground returns, package resonances, and over-moded substrates. As such, we have a dedicated support team with decades of Sonnet experience to answer questions and help our customers make the most effective use of our products."

Now that we've addressed whether or not Sonnet users need to know Maxwell's equations themselves, another question could be: *Should* Sonnet users know Maxwell's equations? Rautio answered with, "I would suggest that knowing Maxwell's equations covers a broad spectrum. Some of our most dedicated power users are using Maxwell's equations to derive Green's functions, Hankel functions, etc. That knowledge can help those users drive the tool to its absolute maximum potential—even going so far as to script new features."

Rautio continued, "Conversely, we have multidisciplinary users who are interested in validating something specific, and thus have little need to understand the physics beyond successfully extracting a model. In practice, I think there is a "sweet spot," where the median user of our tool benefits from understanding the general physics of the equations—how waves propagate, types of coupling, resonances, and things of that nature—without having to memorize the underlying equations or be ready to evaluate sets of double integrals. It's roughly analogous to the hardware we compute on—knowing how the CPU and RAM interact can be quite important to a user but understanding the VLSI architecture of the CPU is extra credit."

So, there you have some perspective concerning the topic of Maxwell's equations for today's engineers. Of course, only you can determine if you really need to know them or not. But perhaps reading this has motivated you to open up that dusty textbook once again in search of those equations.

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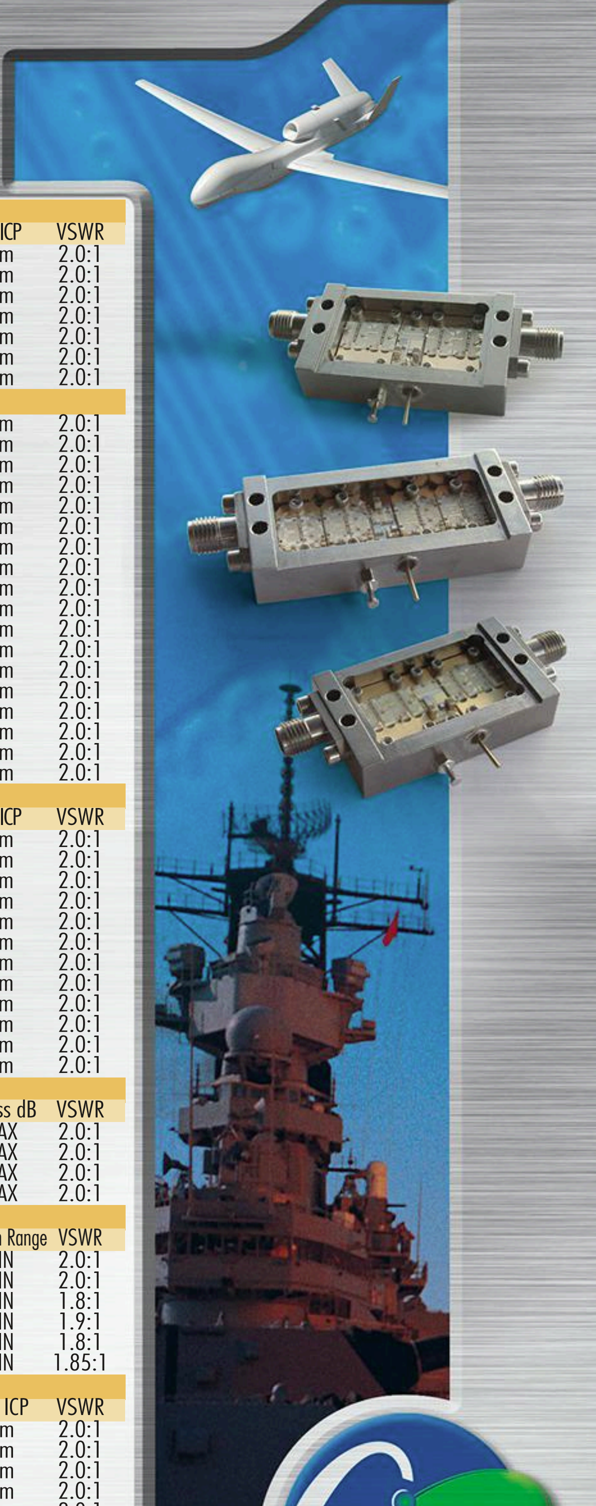

OCTAVE BAND LOW NOISE AMPLIFIERS

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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Long-Standing Company Reveals Many New Tricks at IMS

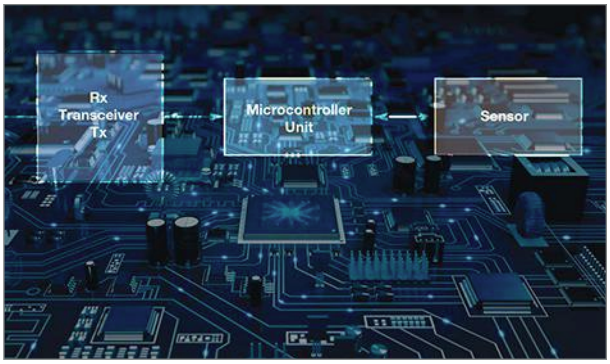
Maury Microwave recently unveiled a host of new solutions, such as updated mixed-signal active load-pull systems and characterized device calibration kits.

https://www.mwrf.com/test-measurement/long-standing-company-reveals-many-new-tricks-ims

How to Pick the Best Bluetooth Protocol for Your Application

From BLE BR/EDR to BLE to Bluetooth 5, the wireless communications technology has gone through numerous variations to meet disparate needs. What exactly are the differences between them?

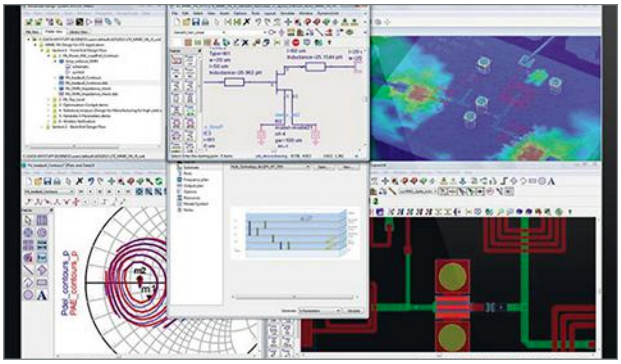
https://www.mwrf.com/systems/how-pick-best-bluetooth-protocol-your-application



A pH Sensor Reference Design Enabled for Wireless Transmission (Part 2)

The second installment of this two-part series on pH wireless sensor monitoring explores topics like hardware design solutions and software implementations.

https://www.mwrf.com/systems/ph-sensor-reference-design-enabled-wireless-transmission-part-2



Take a Path to Accelerated Design Workflows

This software suite is intended to accelerate design workflows for applications across the board, ranging from 5G to automotive radar.

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Where can we take you next?

News

Phase-Noise Test System REACHES THE LIMITS OF PHYSICS

Never one to rest on its laurels, Keysight Technologies *(www.keysight.com)* made a big impression at the recent IMS 2019 thanks to several announcements. One such revelation centered on the launch of the company's new N5511A phase-noise test system (PNTS), which, according to the company, is designed for phase-noise "power users" at the high end of the market (*see figure*).

Keysight emphasizes that the N5511A PNTS can measure phase noise down to the thermal phase-noise floor (–177 dBm/Hz), which is the theoretical limit for any measurement at room temperature. What this means is that the N5511A can measure at the limits of physics.

One of the significant features of the N5511A is its dual-channel cross-

The N5511A phase-noise test system can measure phase noise down to –177 dBm/Hz.

spectral averaging (cross-correlation) capability. In this configuration, the phase noise of the reference does not limit the effective noise floor of the absolute phase-noise measurement. Furthermore, the internal measurement system noise generated within the phase-detector module is suppressed via the cross-correlation process. By suppressing both reference noise and measurement system noise, cross-correlation results in a sensitivity down to –177 dBm/Hz at room temperature, as mentioned.

In terms of the N5511A's frequency range, customers have a few options at their disposal. Option 503 equips the system with a carrier frequency range of 50 kHz to 3 GHz, while option 526 allows for a range of 50 kHz to 26.5 GHz. Option 540 is the highest frequency option—it enables a frequency range of 50 kHz to 40 GHz.

However, it's also possible to perform measurements at carrier frequencies above 40 GHz by employing external components. The N5511A contains the M9550A phase-detector module, which can operate at those carrier frequencies. To make measurements at carrier frequencies above 40 GHz (i.e., beyond 100 GHz), one must use external double-balanced mixers with a 0-Hz IF capability. The external mixer operates as a phase detector and drives the fully calibrated baseband noise input port on the M9550A phase-detector module.

Carrier offset frequencies range from 0.01 Hz to 160 MHz. Such performance is enabled by the M9551A data-converter module inside the N5511A. This module allows for an offset range of 0.01 Hz to 160 MHz thanks to four on-board analog-to-digital converters (ADCs) and a large FPGA for fast FFT processing. Furthermore, the offset frequency range can extend to 3 GHz by using an external signal analyzer. ■

AVX RECEIVES 2018 TTI Asia Supplier Award

AVX CORP. (WWW.AVX.COM) earned a "threepeat," taking home the Supplier Excellence Award from TTI Asia for 2018. It marks the third time that earned this prestigious recognition of outstanding service and delivery to customers throughout Asia (*see figure*). The award recognizes such factors as quality acceptance, ship-to-commit delivery date, effective business systems, ease of doing business, and both the quality and efficacy of field employee and management relationships.

AVX Corp. won the 2018 Supplier Excellence Award, the third time it has been awarded the recognition by TTI Asia. *(Courtesy of AVX Corp.)*

The firm is a global supplier of a wide range of circuit elements and components, including ceramic capacitors, diodes, antennas, filters, and interconnections. It holds many patents and is aiding the growth of emerging high-volume electronic markets, including in 5G wireless cellular networks and equipment and Internet of Things (IoT) devices and components. The award was presented earlier this year by Anthony Chan, president of TTI Asia Pacific, to Alex Schenkel, AVX senior vice-president of sales.

"The 2018 TTI Asia Supplier Excellence Award marks the third time AVX has been recognized in our program, which has been instrumental in elevating supplier and distributor performance since being introduced in Asia in 2007," said Sam Sung, vice-president, Product and Supplier Marketing, TTI Asia. "This is a significant achievement that not only represents the ongoing successful partnership between AVX and TTI Asia, but also the dedication of all the AVX Asia employees to our program and their commitment to providing exceptional quality, on-time delivery, and outstanding customer service."

Schenkel responded, "We are honored to have received our third TTI Asia Supplier Excellence Award, and are very proud to have been recognized for our success by such a valued strategic partner as TTI. AVX consistently strives to exceed expectations with regard to customer satisfaction and is committed to continuously improving upon our quality performance." ■

DIGITIZER BRINGS FUNCTIONALITY AND POWER to the Table

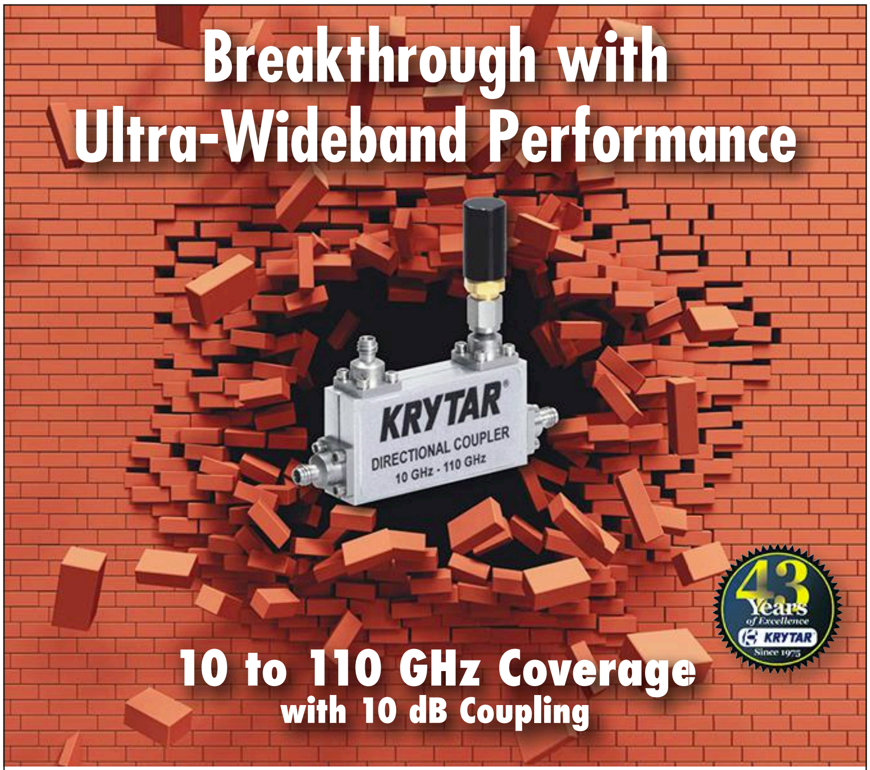
The 6 Series low-profile digitizer is equipped with four channels and offers a sample rate as high as 25 Gsamples/s.

TEKTRONIX (WWW.TEK.COM) has been particularly busy of late bringing new test-and-measurement equipment to the market. One recent unveiling is the 6 Series low-profile digitizer (model number LPD64) *(see figure)*. The company states that this new high-speed digitizer "has the functionality of a digitizer and the power of an oscilloscope," and that its hardware platform is similar to the 6 Series MSO oscilloscopes.

Equipped with four-channels, the 6 Series low-profile digitizer offers as much as 8 GHz of bandwidth and achieves a sample rate as high as 25 Gsamples/s. Supplied in a 2U form factor, the digitizer also features 12-bit analog-to-digital converters (ADCs) and has a standard record length of 125 Mpoints (the record length can be extended to 250 Mpoints as an option). Furthermore, SMA connectors provide the interface for all four channel inputs.

A key component for all channels is a real-time digital downconverter (DDC)—it allows for a capture bandwidth as high as 2 GHz when equipped with an optional analysis capability. Additional options provide customers with features like remote I/Q data transferring and RF versus time analysis.

Looking under the hood of the 6 Series low-profile digitizer reveals the Tektronix-designed TEK049 application-specific integrated circuit (ASIC). This ASIC contains 12-bit ADCs, which provide 16 times more resolution than traditional 8-bit ADCs. The TEK049 is paired with the TEK061 front-end ASIC, which, according to the company, enables "breakthrough" noise performance at the highest sensitivity settings. Thanks to the TEK061, RMS noise is only 54.8 μV at 1 mV/div with a 1-GHz bandwidth. ■



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RFID COULD WEAR WELL in Clothing

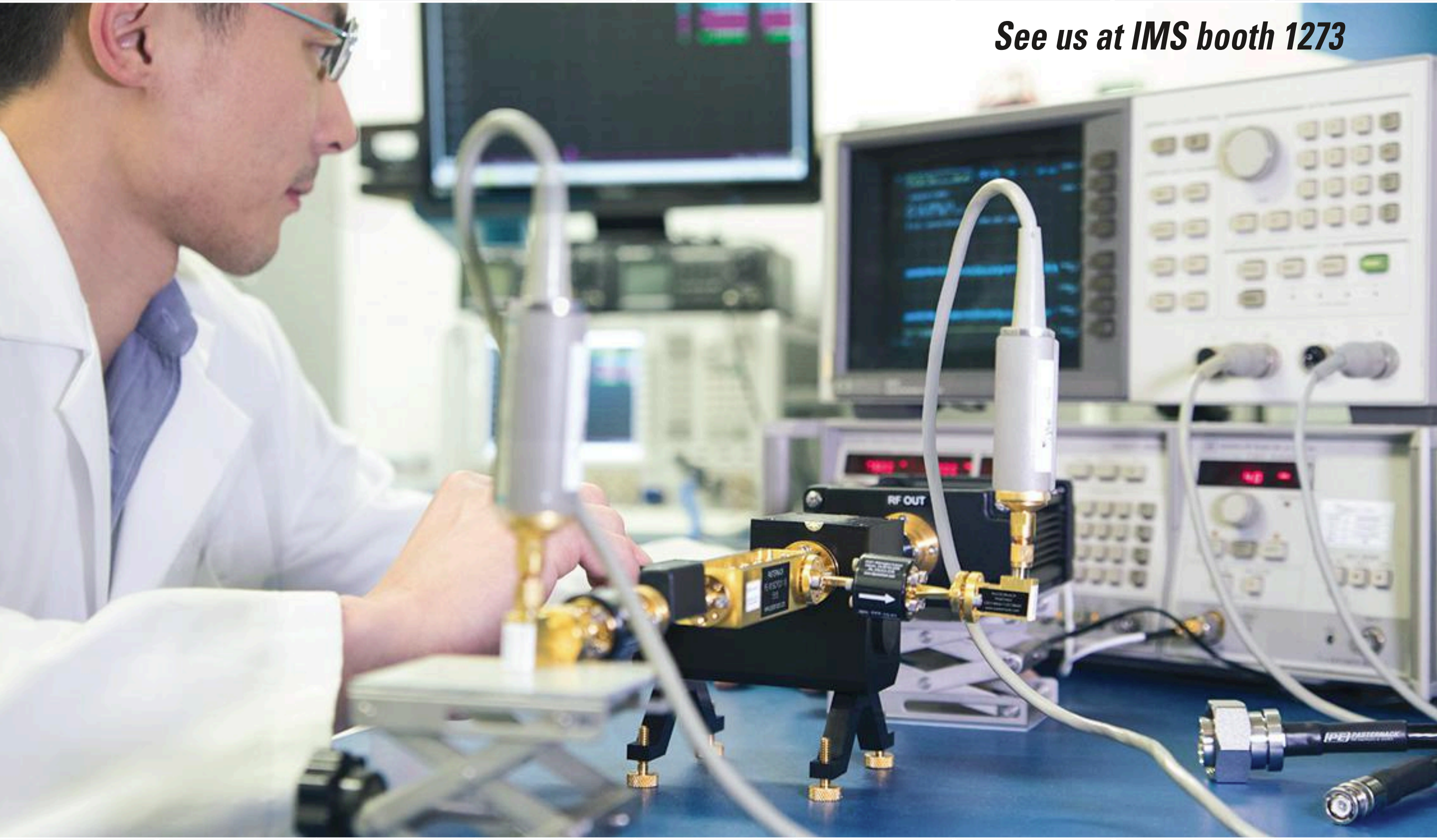
RADIO-FREQUENCY IDENTIFICATION (RFID), is a low-cost wireless technology that provides a practical means of sorting through different items when used as electronic tags. RFID has also been increasingly used to add identification functionality to clothing, such as for electronic ski-lift passes at vacation resorts.

Justin Patton, director of the RFID Lab at Auburn University, feels that there's great potential for growth in the use of RFID technology in clothing. However, conquering counterfeiting will be one of the concerns in establishing the widespread use of RFID as part of the "fabric" of commercial clothing products.

As Patton notes, "Most ski-lift passes are RFID. There is a concept to embed a UHF RFID tag in a ski jacket so that a skier could just enable the lift pass directly to their jacket and not have to carry a separate item. I don't think we've really cracked the killer app for wearable persistent RFID yet, but the capability is definitely there."

One of the concerns for embedded RFID tags in garments will be to properly apply the technology to avoid counterfeiting of clothing products. "Counterfeiting is a massive market, and it's getting worse," explains Patton. With uniquely coded RFID tags embedded in clothing products, customers can be assured of receiving authentic goods for the premium prices they pay. ■

Justin Patton is the director of the RFID Lab at Auburn University, which focuses on the application of emerging technologies for retail, supply chain, manufacturing, and aerospace uses. *(Courtesy of Auburn University)*

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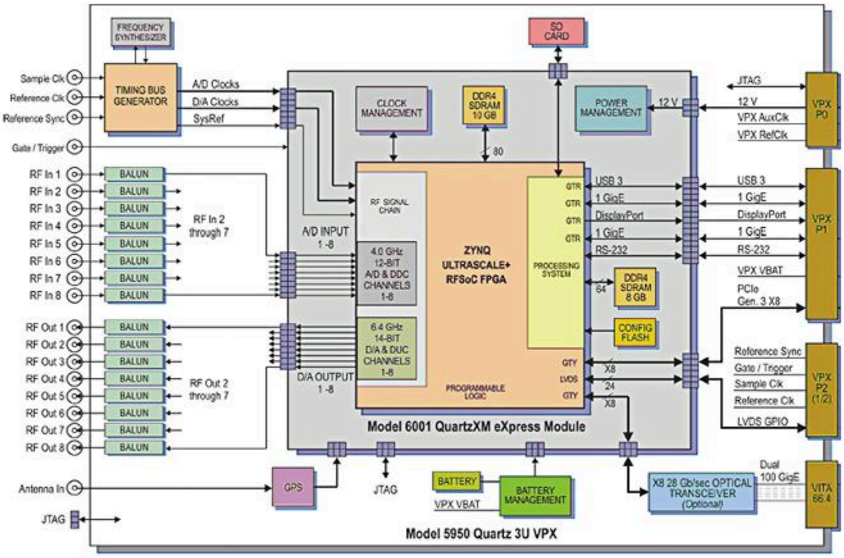
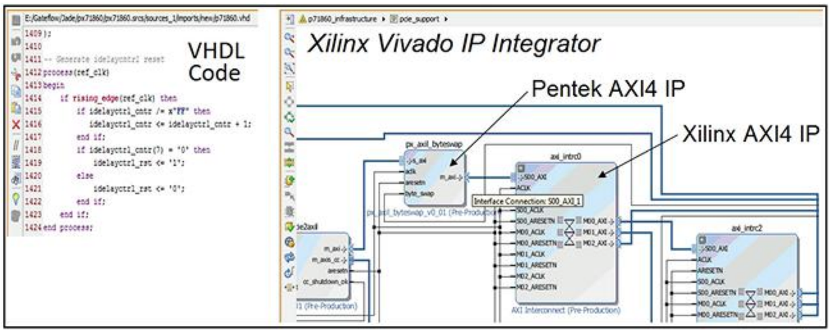
These COTS SDR System Solutions FOCUS ON 5G

Future 5G implementations will require many development platforms for experimentation. Using a COTS system as a starting point will accelerate time to market.



1. Xilinx's RFSoC is the central component of the Pentek 5950 3U VPX board.



```
);

-- Generate idelayctrl reset
process(ref_clk)
begin
    if rising_edge(ref_clk) then
        if idelayctrl_cntr /= x"FF" then
            idelayctrl_cntr <= idelayctrl_cntr + 1;
        end if;
        if idelayctrl_cntr(7) = '0' then
            idelayctrl_rst <= '1';
        else
            idelayctrl_rst <= '0';
        end if;
    end if;
end process;
```

2. The Navigator FPGA Design Kit (FDK) graphical block representation (right) provides better insight than the VHDL code (left).

Commercial-off-the-shelf (COTS) software-defined radio (SDR) products have been traditionally used for military radar and communication applications due to their performance and design flexibility. The latest SDR products offer solutions with integrated I/O, Arm processors, and large FPGAs that include intellectual property (IP) for accessing, routing, and processing digital data. Combining these attributes with superior signal integrity, phase-coherent sampling, and multichannel transceivers, a COTS SDR system becomes an ideal choice for a 5G development platform.

THE LATEST COTS SDR TECHNOLOGY

Over the past 10 years, FPGA manufacturers like Xilinx have been improving technology by reducing the silicon fabrication structure size and, as a result, the device's size, weight, and power (SWaP) values. The latest system-on-chip (SoC) device from Xilinx, the RFSoC, consists of FPGA fabric with Arm processors, analog-to-digital converters (ADCs), and digital-to-analog converters (DACs) all on the same chip.

The 16-nm technology has over 4.2K DSP slices, four 1.5-GHz A53 Arm processors, two 600-MHz R5 ARM processors, eight 4-GHz, 12-bit ADCs, and eight 6.4-GHz, 14-bit DACs per device. This game-changing technology can be used by COTS manufacturers to provide multichannel, SDR transceivers for engineers developing 5G radio products.

Figure 1 shows a functional block diagram of one COTS implementation of the Xilinx RFSoC—it's the central component of the 5950 3U VPX board from Pentek. The "gray" area is a fully connectorized RFSoC or system-on-module (SOM) that plugs into the 3U VPX carrier. This device can be controlled via the Gigabit ENET port similar to the previous-generation FPGA. However, the on-board Arm processors allow for autonomous operation and the ability to communicate with, or control devices locally on, an external network.

Previous-generation FPGAs were programmed using a textual hardware description language, or HDL like Verilog, or VHDL. The latest AXI4-compliant IP blocks are included in Vivado from Xilinx. The IP Integrator tool from Xilinx has virtual graphical blocks that represent HDL code, which can be connected to one another via drag-and-drop wiring. *Figure 2* shows VHDL code on the left in contrast to the graphical block representation on the right.

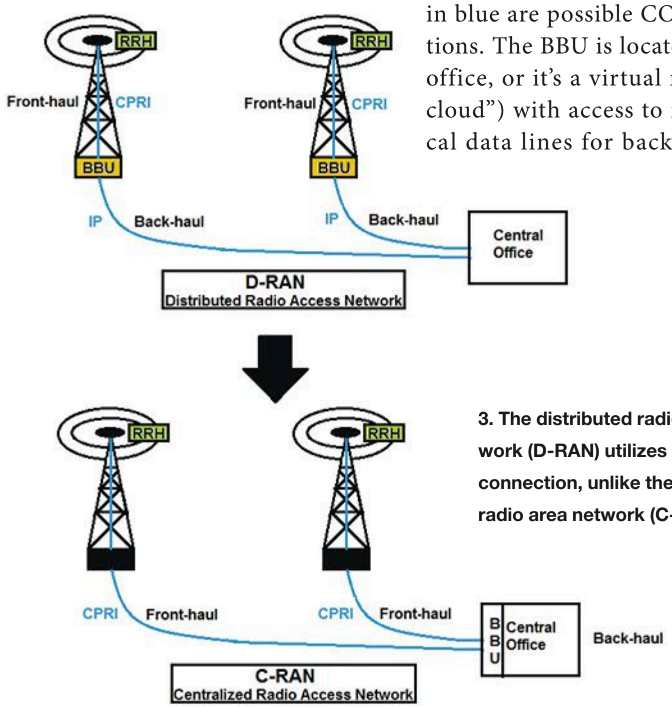
This more intuitive way to program enables someone new to FPGAs to wire together logical blocks that represent hardware like FIR filters and DDCs to create an SDR. Such a programming method supports fast integration of vendor-supplied, hardware-specific IP blocks with Xilinx IP blocks to create a working SDR system. Both IP block types can be combined to create a common library.

These IP programming advances provided an opportunity for COTS vendors to create a single board-support-package (BSP) module that corresponds to one IP module with all of the necessary FPGA program parameters in one location. An example would be a "Clock Control BSP Module" that corresponds directly to a "Clock Control IP Module."

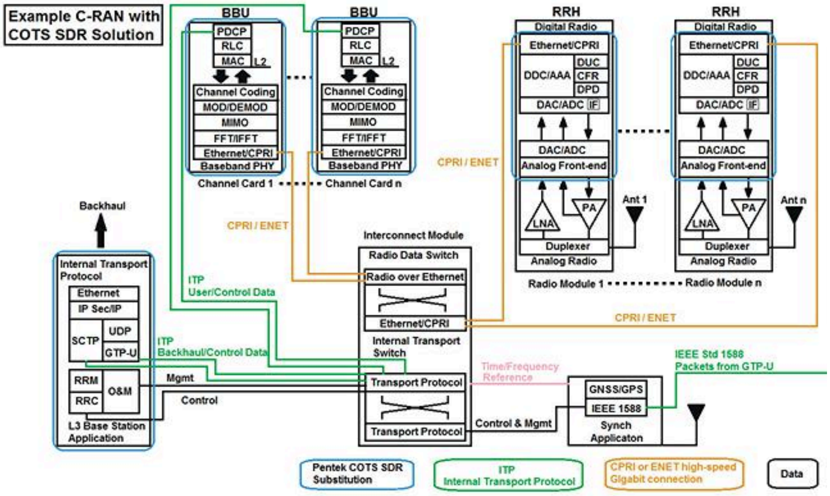
5G APPLICATION-SPECIFIC EXAMPLE

Figure 3 illustrates the difference between a distributed and centralized radio area network (D-RAN and C-RAN). The traditional D-RAN "cell sites" were initially being replaced by newer C-RANs to improve data-transfer efficiency and reduce radio costs, but the latest 5G millimeter-wave massive-MIMO applications require the separation to move the remote radio head (RRH) closer to the end user because of increased RF path loss.

Figure 4 is a functional block diagram of a C-RAN, which consists of a baseband unit (BBU), RRH, GPS time and frequency reference, and an interconnect module. The selections in blue are possible COTS SDR locations. The BBU is located at a central office, or it's a virtual network ("the cloud") with access to multiple optical data lines for backhaul, and the



3. The distributed radio access network (D-RAN) utilizes an IP backhaul connection, unlike the centralized radio area network (C-RAN).



4. A centralized radio area network (C-RAN) consists of a baseband unit (BBU), remote radio head (RRH), GPS time and frequency reference, and an interconnect module.

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NEW!	MGA-718544-HP3	7.1-8.5	13	44	HP3
NEW!	MGA-515844-99	5.1-5.8	16	44	99
	MGA-242740-02	2.4-2.7	15	40	02
	MGA-333840-02	3.3-3.8	12	40	02
	MGA-444940-02	4.4-4.9	12	40	02
	MGA-445343-99	4.4-5.3	14	43	99
	MGA-495940-02	4.9-5.9	16	44	02

GaAs Based Drivers and PAs

	MODEL	FREQ (GHz)	GAIN (dB)	P-1 (dBm)	QFN Pkg.
	MMA-011015-C3	1.0-10	15	16	C3
	MMA-053223-M4	0.5-3.2	12	24	M4
	MMA-020624-M4	2.0-6.0	17	25	M4
	MMA-054025-M4	0.5-4.0	11	25	M4
NEW!	MMA-445933H-M5	4.4-5.9	31	33	M5
	MMA-495933-M5	4.9-5.9	10	33	M5
NEW!	MMA-062020-C3	6.0-20	14	18	C3
NEW!	MMA-012727-M4	0.1-26.5	12	26	M4
NEW!	MMA-012030-M4	0.1-20	12	27	M4
NEW!	MMA-174321-M4	17-43	21	18	M4
NEW!	MMA-243033D-M5	24-30	25	33	M5
NEW!	MMA-212734D-M5	21-27	25	34	M5
NEW!	MMA-273334D-M5	27-33	25	33	M5
NEW!	MMA-004023D-M4	0.03-40	16	23	M4
NEW!	MMA-005022D-M4	0.03-50	15	22	M4

Note: All MMICs parts are also available in chip form.

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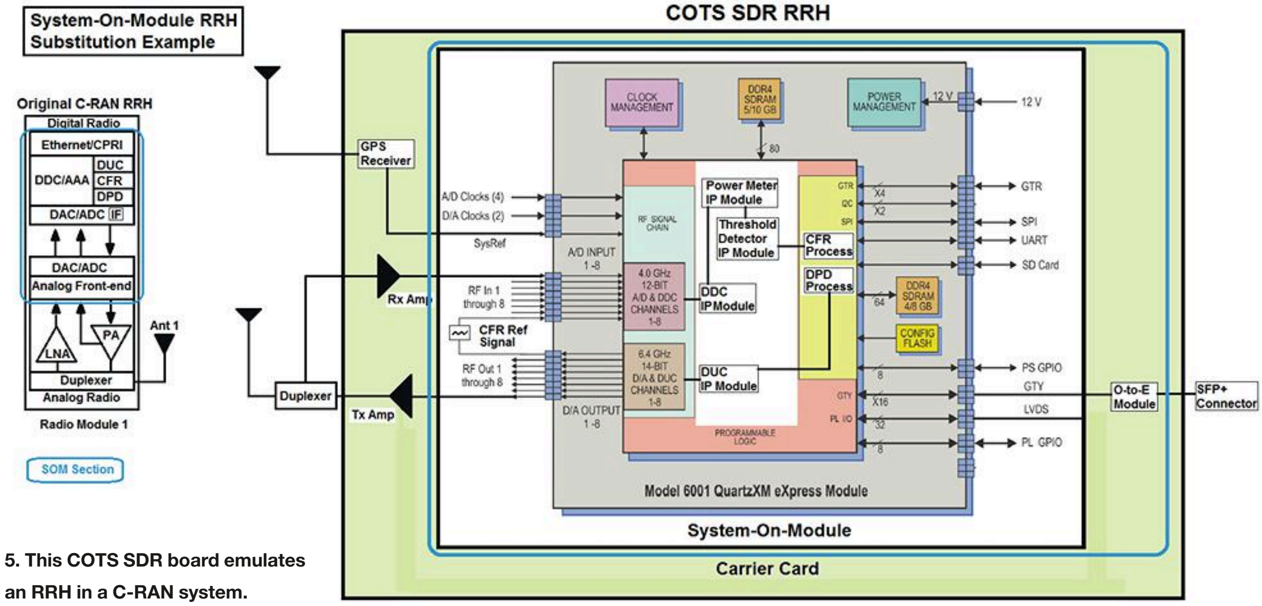
RRH is in an external location closer to the end user. The BBU and RRH in this fronthaul connection example can use a common public radio interface (CPRI), open base-station architecture initiative (OBSAI), or standard Ethernet depending on system requirements. New fronthaul concepts like extensible radio access networks (xRANs) and open radio access networks (oRANs) will be replacing these legacy interfaces in the future.



5. This COTS SDR board emulates an RRH in a C-RAN system.

The various transfer-mode options combined with legacy cellular, 5G TF (Verizon specification), or the 3GPP 5G NR (New Radio) can be configured to form a complex heterogeneous network that will require a flexible development platform.

COTS HARDWARE SOLUTION EXAMPLE

Figure 5 is an example COTS SDR board used to emulate an RRH in a C-RAN system. A subsection of the original C-RAN with the RRH is pictured on the left side and the COTS SDR RRH to the right. The blue encircled areas are equivalent. The custom modular carrier card (light green area) contains Rx and Tx amplifiers, a GPS receiver, and an O/E transceiver module. The SOM (gray area) contains the RFSoC as well as all of the connections for power management, data storage, and analog and digital I/O.

The incoming RF signal from the antenna is connected to the Rx LNA via a duplexer, isolating it from high power-amplifier (PA) transmit levels, and connecting it to one ADC channel. This SOM and custom carrier combination can emulate the original RRH, provided it has the necessary IP described in the next section.

Once inside the FPGA fabric, the digital samples are decimated, frequency-selected or tuned, and filtered in the DDC. The DDC output samples can be streamed to the power-meter module for measurement and sorted in the threshold detector IP module. Subsequently, these processed samples can be streamed to the Arm processors for crest factor reduction and digital predistortion routines before being upconverted in the digital upconverter (DUC) for re-transmission. The DUC is the reverse of the DDC using frequency translation, and interpolation instead of decimation.

The digitized I/Q sample data is packetized in the digital radio for transport to the BBU via a radio data switch, much like that in the previous description of a cellular phone call. Due to the variety of channels and data-transfer protocols, it's necessary to calculate the maximum data throughput of your signal.

Depending on the desired level of control, either BSP routines need to be created for the new IP and Arm processors, or the Arm processors in conjunction with the FPGA can be programmed to operate autonomously.

Such SDR platforms provide superior signal-integrity performance, high test repeatability, and modular assemblies that adjust to constantly changing 5G design requirements. Future 5G implementations will require many development platforms for experimentation. By using a COTS system as a starting point, it will more than likely accelerate the time to market.

BOB MURO, an Application Specialist at Pentek Inc., has over 25 years of work experience in test & measurement engineering from Anritsu, the Wireless Telecom Group (WTG), and LeCroy Oscilloscopes. He holds a BSES & MSBME from the New Jersey Institute of Technology.

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RFoF SIGNALS:

Unlimited Capabilities for the Connected Battlefield

Transmitting RF over fiber allows for greater bandwidth and speed for mission-critical data processing.

High bandwidth and speed are critical factors for network success.

The number of connected devices around the world is growing exponentially—25 billion of them are predicted by 2025, according to research company Gartner. The number of devices connected within the defense industry is also growing rapidly with diagnostic and situational awareness platforms all becoming smarter. This growth is bringing with it the ability to improve military effectiveness through greater connectivity and access to information and intelligence.

As a result, the integration of C4ISR (Command, Control, Communications, Computer, Intelligence, Surveillance, and Reconnaissance) applications is providing unprecedented advances in military capabilities. Critical to the performance of these applications is the communications network, which is relied on to provide secure, reliable, and robust connectivity that does not fail mission-critical communications.

THE CONNECTED BATTLEFIELD

Thanks to the integration of C4ISR applications, the military has gained a competitive advantage with improved decision-making and allocation of resources. Integral to achieving this is the process of data transference between the command, communication, and control systems to collect the information needed from a variety of sources. The data collected enables the military to observe, decide, and act effectively. Any limitation in connectivity will reduce the quality and timeliness of the data provided as well as the quality of the decision-making process. The performance of network connectivity is critical to the C4ISR system and making mission-critical decisions.

CRITICAL CONNECTIVITY

With massive amounts of information passing across the network to power systems and operations, it's critical for the network to deliver high bandwidth and speed capabilities. Any interruption or delay in connection weakens the system and the critical performance of the military. To facilitate data sharing, many entities and systems must be integrated and connected to advance intelligence and operational planning capabilities to deliver an advantageous connected battlefield *(see figure)*.

The geographic areas of the battlefield are large and can span a whole nation. Therefore, the data must efficiently transmit across far distances in real time for command and control systems to execute a perfectly timed decision. As the core architectures remain RF (i.e., antenna and receiver), the demand for an appropriate convergence of technologies has never been greater.

CONNECTIVITY WITHOUT LIMITS

Transmitting RF over fiber (RFoF) allows the defense industry to expand the capacity and speed of its network beyond current constraints in the various theatres. In RFoF systems, electro-optical (EO) converters are used to convert RF to optical at the signal source. This signal is then transmitted across single-mode optical fiber to the destination where optical-electro (OE) converters convert it back into RF.

Key to this conversion is that the inherent analog characteristics of the RF signal are not altered during the conversion process. The result is that RFoF can provide greater bandwidth and speed for mission-critical data processing.

WINNING WITH RFoF

The main advantages of using optical fiber include extremely low transmission losses and immunity to electromagnetic, radio, or other types of signal interference essential for military operations. It's also integral for the military to have unlimited bandwidth at their disposal. Optical fiber allows the signals to span much greater distances integral for battlefields that span vast geographical locations. Since optical fibers are light in weight, they also support weight-sensitive and rapid deployment-focused applications.

Another vital factor in the performance of C4ISR applications is robust network security. Optical fibers feature a high level of security from the antenna to the receiver end, making them much more difficult to tap. With optical fiber, any disturbance is much easier to detect, locate, and prevent from becoming a security threat, which is a priority for mission-critical communication.

In addition, optical fibers can transfer any frequency, any data rate, and/or any modulation format both today and in the future, providing the military with a network they can use to distribute different signal types through the same fiber infrastructure. This creates a flexible network that can be adapted to future military commands. The same infrastructure can also be used without any need for an upgrade, saving time and money. Overall, RFoF is able to provide the military with improved security, performance, readiness, and resilience at lower overall costs.

(Continued on page 79)

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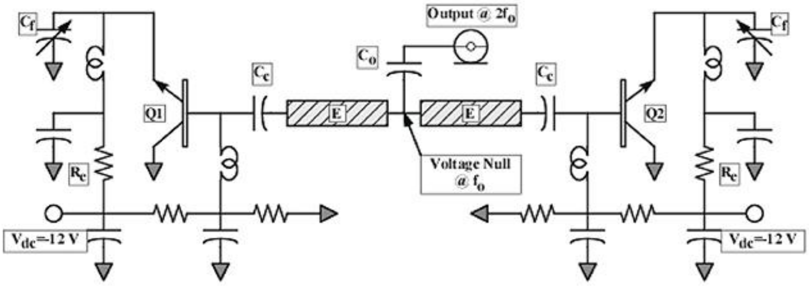
X-band Push-Push Oscillator Simulation and Measurement (Part 2)

Wrapping up this two-part series, we break down the steps involved in simulating and measuring push-push oscillators in terms of output power and voltage tuning.

This article continues where Part 1 left off with a discussion of the push-push oscillator topology. The push-push oscillator circuit topology joins two common-collector oscillators at the grounded end of the resonator *(Fig. 1)*. The voltage null of each section of the oscillator becomes the common junction of the push-push configuration, and the coupled transmission line is removed in favor of a simple coupling capacitor.



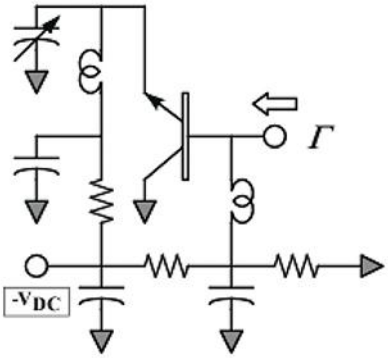
1. This is the push-push oscillator circuit topology. The push-push oscillator is structured from the connection of two common-collector oscillators such that the null point of the circuit is maintained.

For a voltage null to exist at the junction of the oscillator sections, each oscillator must generate a signal that's coherent and in phase opposition. Clearly, symmetry and circuit balance are requirements for quality push-push oscillator operation. The second harmonic of the oscillator sections is extracted at the null point while the fundamental is cancelled. The primary frequency acts as an idler to produce an output signal with significant second-harmonic content. This is accomplished principally because the fundamental signal component isn't loaded. Therefore, highly nonlinear operation is assured.

Although the grounded collector oscillator has been selected, it should be emphasized that other oscillator configurations are applicable—in fact, other types of devices are applicable. The principal criterion is that a negative resistance is required, as is a means of joining the negative resistance devices in a manner that ensures synchronization and out-of-phase signal generation from each device.

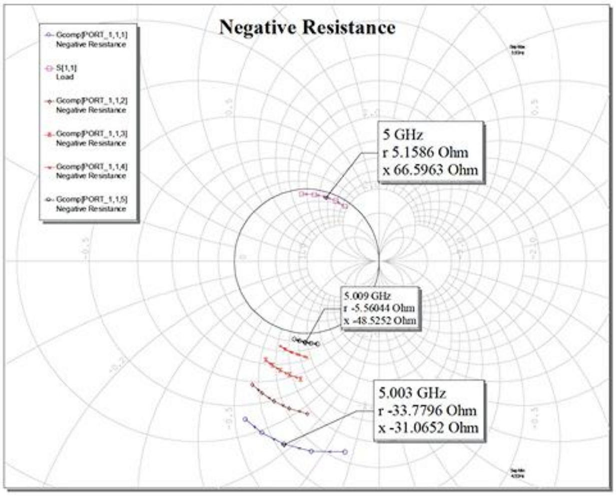
SIMULATION TECHNIQUE

The bipolar junction transistor (AT42000) is initially investigated under small- and large-signal conditions using the feedback capacitor to maximize the reflection coefficient over the anticipated bandwidth of the oscillator. This investigation is conducted using the schematic shown in *Figure 2*, in which the device is biased at a nominal operating point and a signal is applied to the negative resistance port at variable power levels.



2. Initial negative resistance investigation is conducted using this schematic. The reflection coefficient is optimized with the feedback capacitor and plotted as the input power is increased.

After optimization of the reflection coefficient at the base under small-signal conditions using the variable feedback capacitor, the reflection coefficient is repeatedly measured under increasing input RF power conditions.



3. Large-signal reflection coefficient is compared with drive power. Note that the large-signal reflection coefficient decreases as input power is increased. This observation validates oscillator theory with respect to the buildup of signal to the steady-state limiting—the net oscillator loop resistance is zero at steady state.

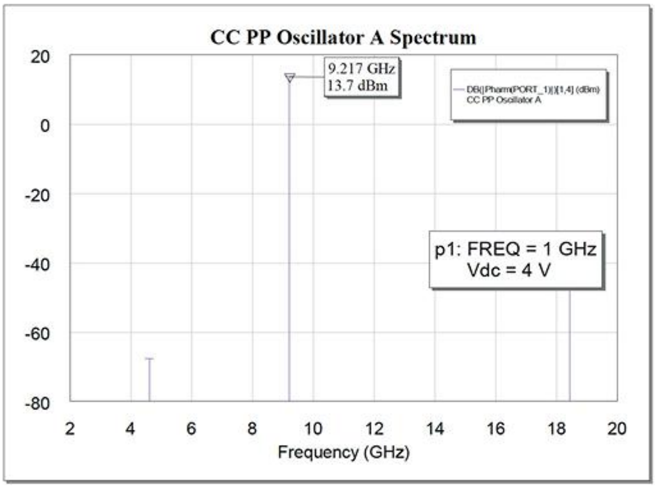
As may be readily discerned from *Figure 3*, the reflection coefficient decreases as the signal level increases, as one might expect from an intuitive understanding regarding the signal limiting of oscillators. Note that the negative resistance decreases with increasing RF input power. This is expected and mirrors the phenomenon that occurs as the oscillator starts and settles to the final operational conditions.

Such an exercise may also be used to predict oscillation frequency—for example, continue to increase the signal level and observe the point where the reflection coefficient magnitude is equal to unity. At this point, one may read the real and imaginary parts of the impedance. The reflected impedance from the coupled resonator is also illustrated. The point at which the negative resistance at the base of the transistor is equal to the coupled port positive resistance and the total loop reactance cancels is the oscillation frequency.

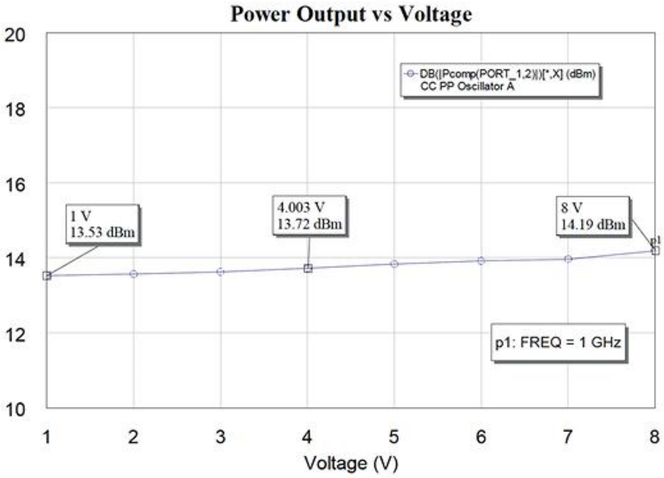
This is an estimate and most large-signal simulation algorithms are much more sophisticated in both technique and execution. However, it's instructive to observe the behavior and validate the traditional understanding of oscillator theory.

It's also instructive at this point to alter the feedback capacitance to ensure that the maximum reflection coefficient is maintained at the base of the transistor. To ensure that the oscillator starts under small-signal conditions is the first requisite; to maintain a healthy oscillator, the feedback capacitor must also sustain the large-signal limiting conditions. A phenomenon that's been observed is the oscillation cycle of start, saturation, and subsequent quenching.

The common-collector oscillator was further explored using the coupled transmission line as a parametric variable.



4. The spectral content at Vt = 4.0 V dc shows the strong second harmonic as expected with fundamental, third-, and fourth-harmonic levels present.



5. Shown is power output versus tune voltage. The oscillator power output remains constant within 0.25 dB over a voltage tuning range of 8 V.

The parameters of output power, phase noise, and load line were documented for various coupling values.

SIMULATION RESULTS

The simulation algorithm begins with a frequency search of the required conditions for oscillation under small-signal conditions. It proceeds to find a large-signal, steady-state solution at the identified spectral areas using a probe element as described in *Appendix B. Figures 4, 5, 6, 7, 8,* and *9* reveal the simulation data.

J.M. Ziman once said, "It is typical of modern physicists that they will erect skyscrapers of theory upon the slender foundations of outrageously simplified models." Like the modern

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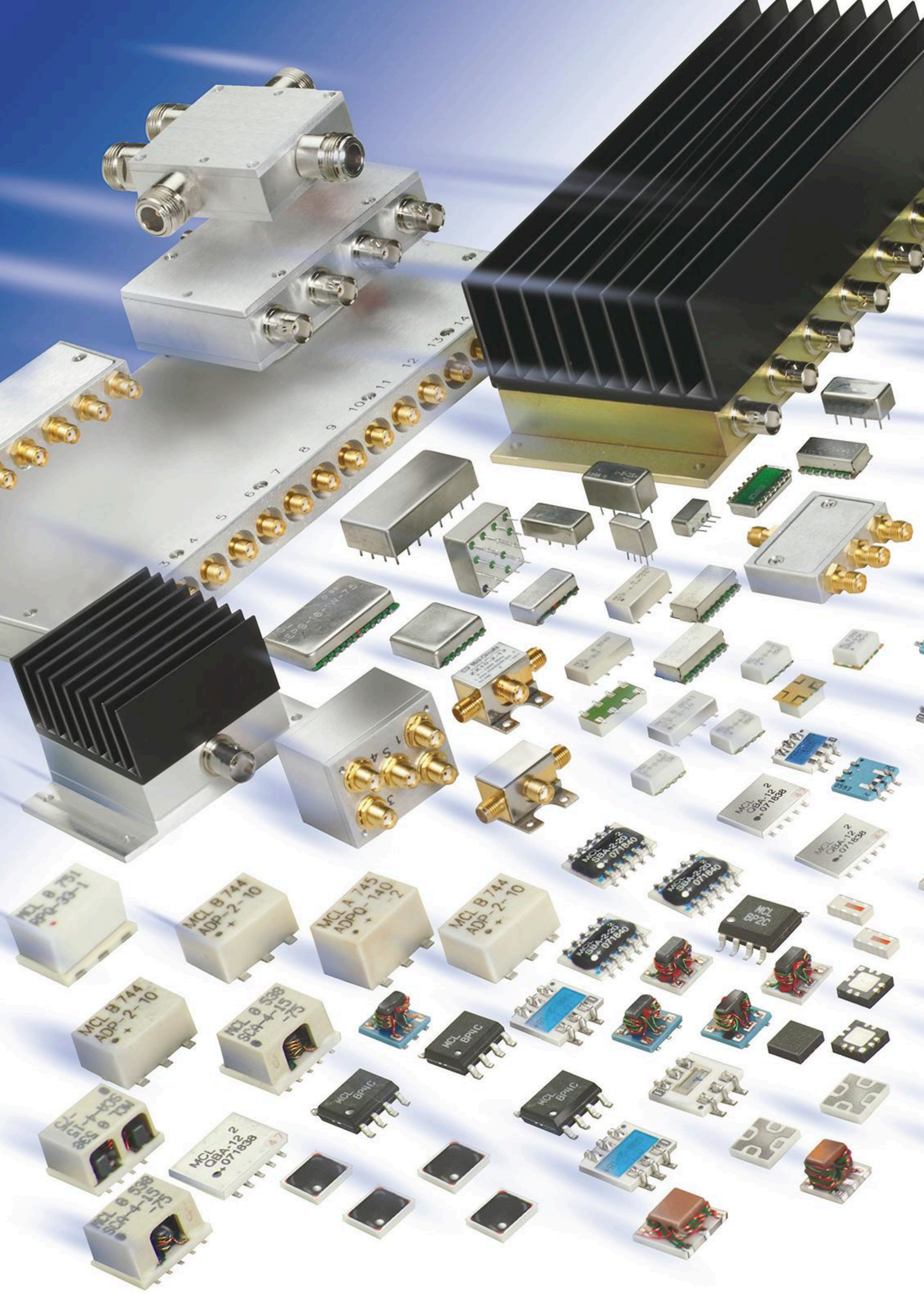
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physicist, the engineer should be cautious in the use of overly simplified models.

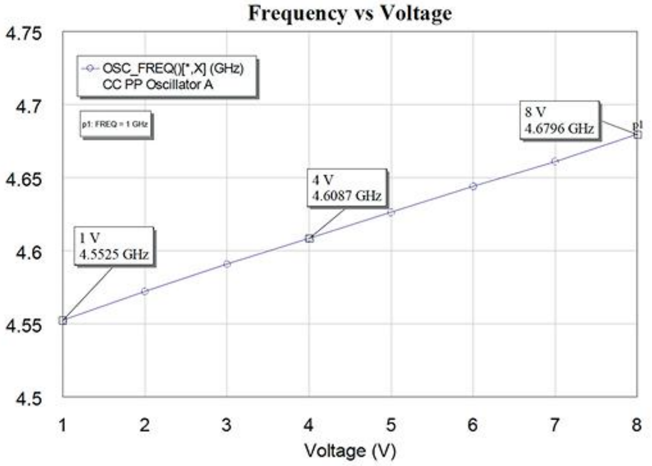
PREDICTION OF OSCILLATOR CENTER FREQUENCY

The simulation utilized a published Spice model of the AT420 bipolar transistor. Attempts to validate the model from the manufacturer were not successful. In fact, S-parameter data did not correlate well between the manufacturer's published data and S-parameters simulated with the Spice model. This lack of correlation is believed to be responsible for the poor correlation between the measured (10 GHz) and simulated center frequency (9.2 GHz).

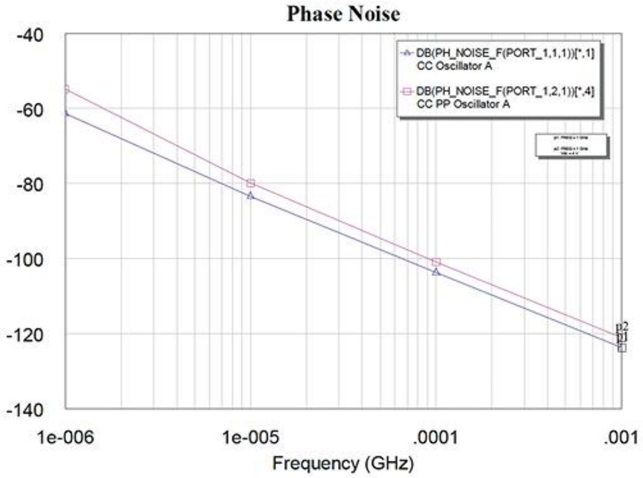
PREDICTION OF OSCILLATOR PHASE NOISE

The measured oscillator phase noise was −45, −73, and −100 dBc/Hz at the respective offset frequencies of 1, 10, and 100 kHz. The simulated phase-noise data is −61, −84, and −104 dBc/Hz at the respective offset frequencies. Phase noise of a free-running oscillator near to the carrier is closely related to the Spice model noise parameters. Unfortunately, Spice noise parameters are generally not published and when published, disclaimers sometimes follow with respect to production lot variations and measurement methods.

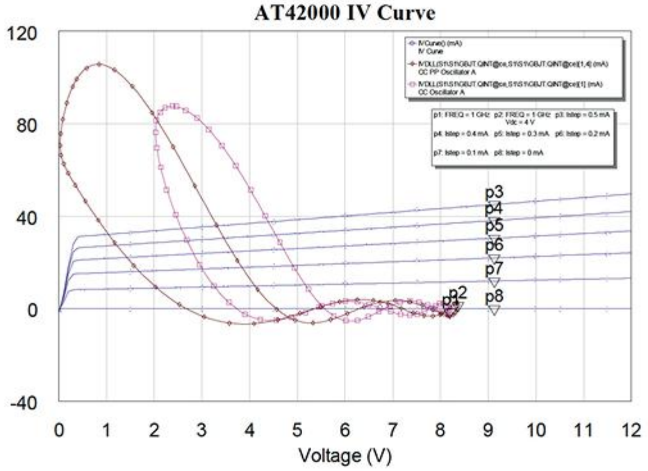
The Spice noise parameters—AF, KF, and FFE—for the AT420 transistor are not available from the manufacturer.



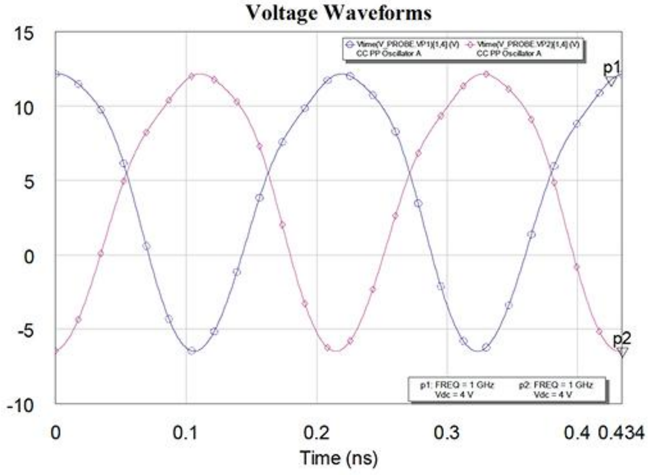
6. The oscillator output frequency (fundamental) versus tuning voltage discloses 250 MHz of voltage tuning over 8.0 V. The tuning range is a strong function of the varactor coupling capacitors. A tuning range of 1.0 GHz has been noted with larger coupling capacitors.



7. Shown is phase noise versus offset frequency of the push-push oscillator and the single common-collector oscillator. Because one expects a 6-dB difference, some noise suppression is indicated. Phase noise of −100 dBc/Hz has been measured on the push-push configuration, indicating excellent correlation.



8. The dynamic load line of the push-push oscillator and the single common-collector oscillator discloses profound nonlinear operation of the push-push oscillator. Operation at high collector current and low collector voltage has been correlated with poor phase noise.

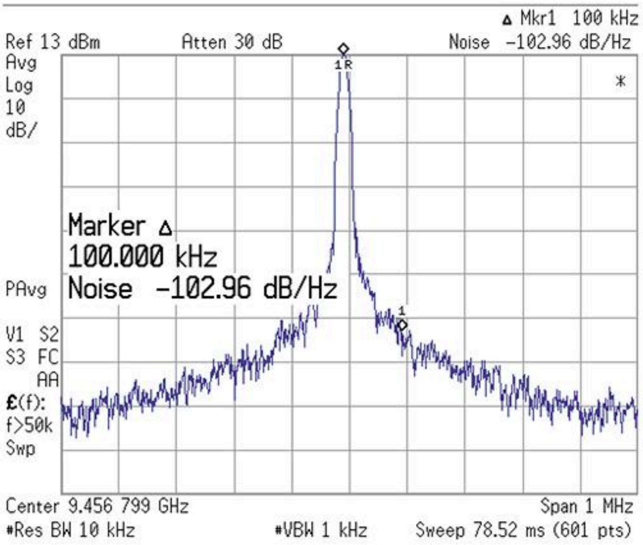


9. Shown are the voltage waveforms at the base of the push-push stages, which disclose the out-of-phase signals as expected.

> "It is typical of modern physicists that they will erect skyscrapers of theory upon the slender foundations of outrageously simplified models."—J.M. Ziman

Therefore, the author relied on measured data from Gris.[1] The deviation between measured and simulated phase noise was not unexpected.

Measurements of a second push-push oscillator disclosed better correlation *(Fig. 10)*.



10. This is the phase noise of the prototype push-push oscillator.

PREDICTION OF OSCILLATOR OUTPUT POWER

The measured oscillator output power included a saturated buffer amplifier. Therefore, correlation between simulated output power could not be immediately determined.

PREDICTION OF OSCILLATOR VOLTAGE TUNING

The oscillator frequency versus voltage tuning was 250 MHz (simulated) compared to 575 MHz (measured) for the push-push prototype. Clearly, further investigation is required. The oscillator tuning versus voltage relies on the varactor model and the coupling capacitor. The hyper-abrupt tuning varactor part number is MHV514-21, from Micro-

metrics. Micrometrics supplied the Gamma and capacitance versus voltage data.

The data was entered into the simulation as discrete points and a polynomial regression was used within the program to approximate a continuous function. An error tolerance may be specified to indicate if the polynomial deviates from the discrete point data by more than a specified percent. In this case, the error was less than 2%.

OSCILLATOR SIMULATION ISSUES

Simulation of RF/microwave oscillators presents several limitations with respect to accuracy, particularly under conditions in which the active device operates over a wide dynamic range of current and/or voltage. More specifically, bipolar-junction- and field-effect-transistor (BJT and FET, respectively) noise parameters vary dramatically with bias, and unfortunately, are usually extracted under "nominal" bias conditions.

While parametric extraction at "nominal" bias conditions is quite adequate for small-signal and some large-signal amplifier simulations, it's entirely inadequate for accurate oscillator simulation. An additional consideration is the statistical variation of parametric device data typically associated with manufacturing processes and other tangential considerations, i.e., environment.

Notwithstanding the benefits and improvements in harmonic-balance simulation algorithms, accurate oscillator simulation requires parametric data of the active device over the entire range of operational current and voltage. The extraction of a single set of parameters to represent the active device under static and dynamic operation is a formidable task.

From experience with the push-push oscillator of this investigation and other less-complex oscillator circuit topologies, the fundamental limitations in simulation accuracy are summarized:

- Inaccurate BJT and FET device model parameters and parameter variation; particularly noise parameters
- Resistive and reactive loading of the fundamental and harmonics
- Highly nonlinear device operation, i.e., see the dynamic load-line excursion
- Device manufacturing variance

In addition to accurate model data over large regions of the operating voltage and current, i.e., cutoff to saturation, harmonic-balance algorithms exhibit a profound impact in oscillator simulation performance.

CONCLUSIONS

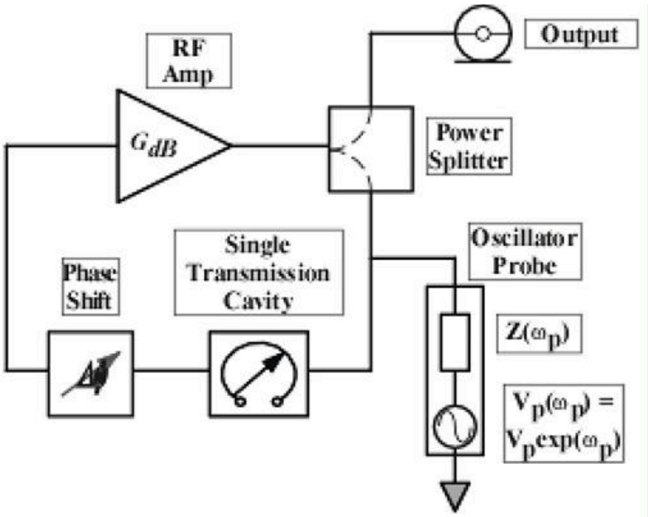
Conventional oscillator theory of operation has been demonstrated using a simple common-collector oscillator configuration. The common-collector oscillator has been extended and shown to be a basic element of the push-push oscillator circuit topology. Highly nonlinear operation, model device parameter extraction, and simulation algorithm are fundamental limitations to accurate oscillator simulation. Notwithstanding accuracy limitations, computer simulation is shown to be an effective educational tool.

THE 2019 PERSPECTIVE

As mentioned in Part 1, this technical documentation was originally written in 2005 to address oscillator tuning issues within a radar subsystem. The exercise created a better understanding of the push-push oscillator circuit topology, and to that extent, the simulation software provided a valuable contribution, notwithstanding the BJT device model limitations.

In retrospect, several elements merit consideration:

- The active device model remains a significant issue with respect to simulation accuracy, particularly phase-noise prediction.
- The Spice model selection remains an appropriate choice for oscillator simulation.
- Discrete BJT and FET oscillator design is becoming—or has become—the province of MMIC manufacturers. Over the last 15 years, the author's oscillator design activities have entailed MMIC device selection, i.e., VCOs, PLLs, and transmit/receive subsystems.
- The unique low-phase-noise property of the push-push oscillator circuit topology is likely the higher quality factor (Q) due to reduced loading at the fundamental frequency, as suggested by Gris[1] and coupled oscillator properties as described in Chang[5] (the Chang reference has been added to the 2019 revision).



A-1. This is a feedback oscillator with probe element.

Simulation of RF/microwave oscillators presents several limitations with respect to accuracy, particularly under conditions in which the active device operates over a wide dynamic range of current and/or voltage.

- The dynamic load line is an effective indication of the degree of nonlinear operation and may be useful in assessing negative resistance and resonator loading.
- The simulation utilized the same common-collector elements. An interesting investigation would encompass intentional asymmetry to disclose performance variations.
- An interesting investigation—and one that the author intends to explore—is the addition of devices to the push-push topology. The performance expectation is increased power output and lower phase noise.

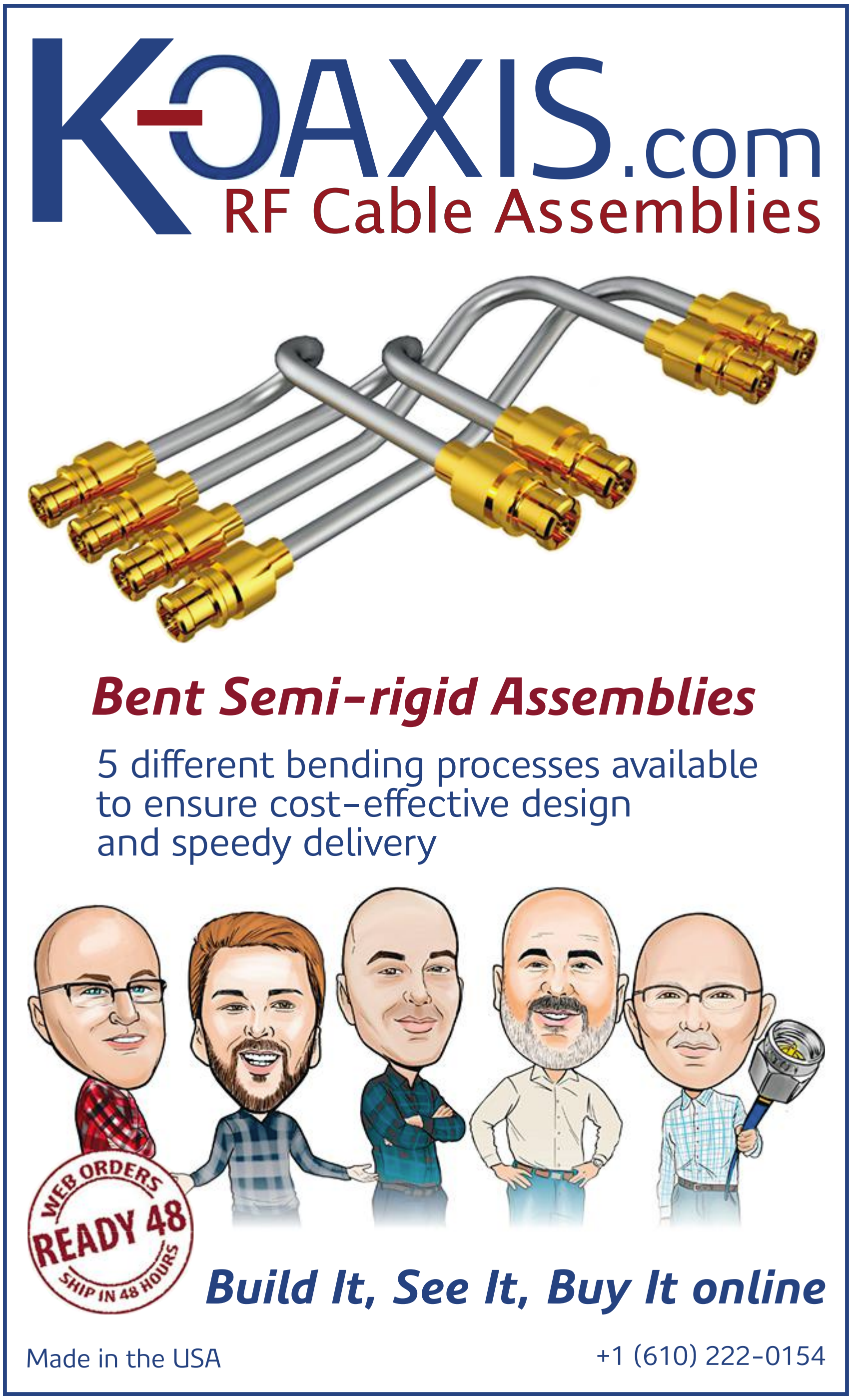
APPENDIX A

Oscillator Simulation Using the Probe Element *(see footnote 1 at end of article)*

The oscillator probe element is a popular technique used in large-signal simulation of oscillators. To understand the operation and simulation algorithm, consider *Figure A-1* in which a typical feedback oscillator is shown with the probe element connected to a resonator located within the feedback loop.

The probe element has the following properties:

- Presents a short circuit at the source frequency and an open circuit elsewhere.
- Probe voltage and frequency are adjusted to exactly equal the steady-state operating voltage.
- No current flows through the probe at the probe frequency or any other harmonic.



The author has found rapid convergence and frequency prediction when the probe element is connected to the frequency-determining element of the oscillator circuit. While this has not been analytically verified, it's suspected that the slope of the reactive impedance at the node of the probe is a component of the search algorithm.

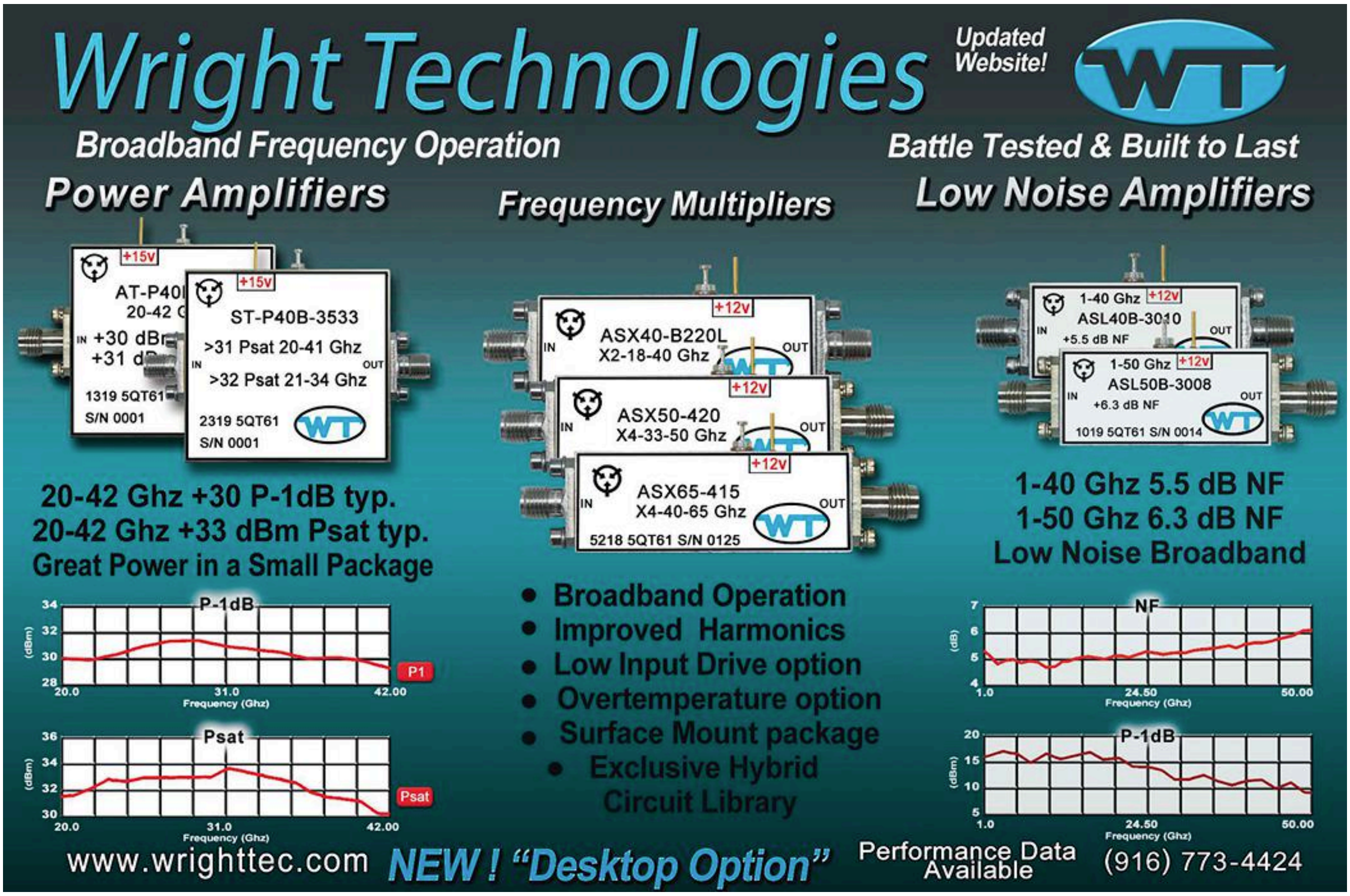
The probe frequency and amplitude are scanned in a search algorithm that identifies the signal frequency and amplitude at the connection point of the probe. When the signal frequency and amplitude are identified at the connection node, the probe no longer disturbs the circuit. Also, its frequency equals the oscillation frequency while its amplitude equals the amplitude at the node to which the oscillator probe is connected. Oscillator analysis is reduced to standard harmonic-balance analysis running in the inner loop of a routine that attempts to locate probe parameters (amplitude and frequency), which result in zero current flow through the probe's terminals.

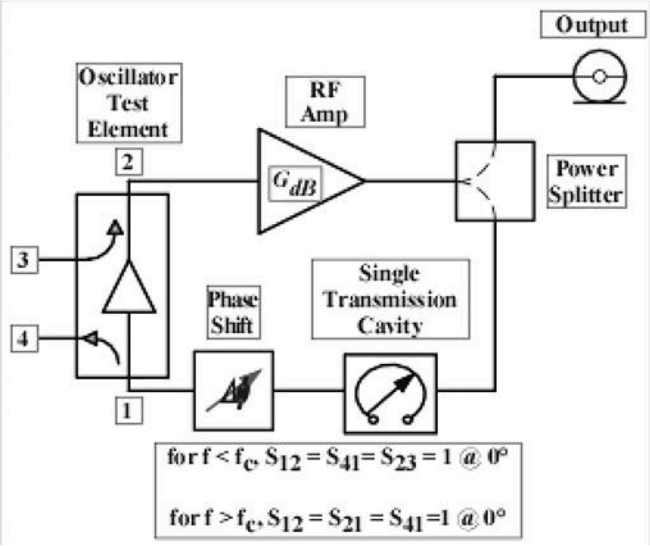
In negative-resistance oscillators, the probe should be connected between the resonator and the negative resistance-generating active device. In feedback oscillators, the probe should be connected to one of the nodes belonging to the feedback loop. Other connections of the oscillator probe generally result in excessive time to achieve convergence of the harmonic simulation algorithm as well as decrease accuracy. Sometimes, an alternate connection will fail to predict oscillation in an otherwise properly operating circuit.

The author has found rapid convergence and frequency prediction when the probe element is connected to the frequency-determining element of the oscillator circuit. While this has not been analytically verified, it's suspected that the slope of the reactive impedance at the node of the probe is a component of the search algorithm.

APPENDIX B

Oscillator Simulation Using the Test Element *(see footnote 2 at end of article)*





B-1. This diagram depicts a feedback oscillator with test element.

The oscillator test element is used to determine loop gain in oscillator design under large-signal conditions.

Referring to *Figure B-1*, large-signal excitation is applied to port 3. The large-signal S-parameter, S_{21}, at the fundamental frequency is monitored. The cutoff frequency, f_c, must be set to any value between the fundamental frequency and the second harmonic. When the excitation is weak, $|S_{21}|$ must be greater than unity to ensure that the start conditions are satisfied. As the excitation level is increased, $|S_{21}|$ decreases as the circuit enters saturation.

At some frequency and some excitation level, $|S_{21}| = 1$ and its phase is zero. This point corresponds to the oscillation frequency. The output power under these conditions is the oscillator's output power. This operational procedure is and has been experimentally verified.

Before the oscillator test element can be used under large-signal conditions, it's best to make an initial small-signal design of the oscillator. The circuit should be adjusted so that $|S_{21}| > 1$ and the phase is zero under small-signal conditions at the desired frequency of oscillation. Increasing the excitation changes the frequency of zero phase shift, and thus it may be necessary to "tweak" the design to keep it on the right frequency. One would expect this phenomenon simply because the active device parameters change with signal level.

The oscillator test element permits the total loop gain to be measured under small- and large-signal conditions and evaluated without disturbing the circuit operation. It's especially useful in analysis of oscillators with discrete feedback loops. An additional feature of the oscillator test element is the determination of frequency control using the phase-shift element. This is accomplished via measurement of large-signal magnitude of S_{21} while varying the phase shift. The oscillator tuning range is determined by observing the frequency points at which large-signal S_{21} is less than unity.

$$[S] = \begin{vmatrix} 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{vmatrix} \quad @ f < f_c$$

and

$$[S] = \begin{vmatrix} 0 & 1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{vmatrix} \quad @ f > f_c$$

B-2. The oscillator test element S-parameters resemble that of a three-port isolator.

The S-parameters of the oscillator test element *(Fig. B-2)* are similar to those of a three-port isolator. They permit circuit excitation without altering circuit function for the measurement of closed-loop gain and phase.

Valid large-signal models of active devices are essential for accurate oscillator simulation. mw

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FOOTNOTES

1. The oscillator probe element, which initiates a large-signal oscillator simulation, is an ideal source in series with an ideal impedance element. The impedance presents an open circuit at all frequencies other than the fundamental frequency of oscillation. Source: Help resource, AWR Design Environment v14.03 Edition.

2. The oscillator test element is used to determine loop gain in oscillator design and break the feedback loop of an oscillator in the forward direction at the fundamental frequency. When this is done, a source at port 3 is used to replace the feedback signal and the feedback itself is measured at port 4. As a result, one can determine the open-loop gain of the circuit under large-signal conditions. Source: Help resource, AWR Design Environment v14.03 Edition.

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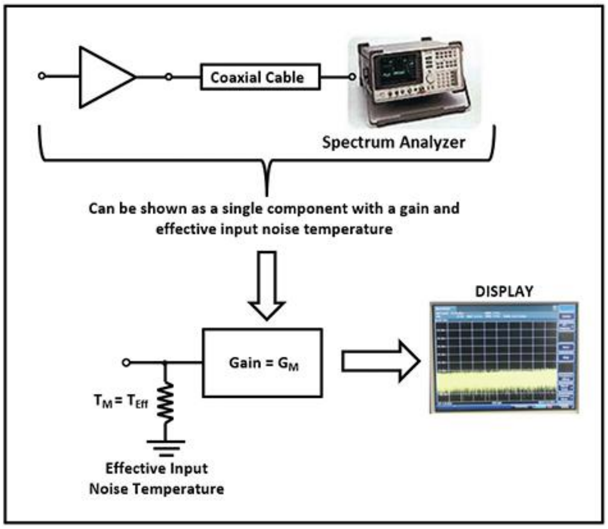
How to Improve Noise-Figure Measurements and Calibrate High-Power Noise Sources

This new technique uses multiple sources to measure any noise source with the accuracy of a calibrated noise-source reference and eliminates the corruption of noise-figure calculations.

Those who have ever tried to manually measure an amplifier's noise figure may have found that the numbers aren't quite what was expected. This is most likely the result of the spectrum analyzer's noise figure, which adds noise to the measurement. If an interfacing amplifier is needed to boost the device-under-test (DUT) output noise power above that of the spectrum analyzer's noise floor, then it becomes even more likely that accuracy will suffer.

Additional components are sometimes necessary, but to accurately determine the noise figure of the DUT, one must know the gain and noise figure of the additional components so that the DUT characteristics can be de-embedded from the measured response. Although high-gain amplifiers are less susceptible to the analyzer's noise figure, some analyzers can have extremely high noise figures at high frequency—on the order of 30 to 60 dB.



1. This is a representation of the measurement test circuit.

In addition to measuring noise figures, sometimes a calibrated high-power noise source is needed when the noise source is built into a system that contains substantial insertion loss in the RF path. This insertion loss is caused by some combination of couplers, splitters, filters, and/or lengths of coaxial line.

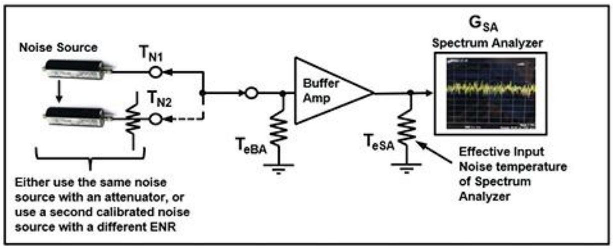
If one could build a perfect RF noise-power measurement instrument that contributed no additional noise to the input noise to be measured, then the two tasks described above would be accomplished. A noiseless measurement instrument is not yet available to the general public.

However, the next best solution does exist: an everyday spectrum analyzer, a calibrated noise source, and a couple of algebraic equations. With these tools, one can measure an unknown noise source within the accuracy of a calibrated reference. This article reveals how to construct this almost ideal test instrument.

BASICS OF THE IDEAL MEASUREMENT INSTRUMENT

The hardware consists of a preamplifier, interconnecting cable, and spectrum analyzer. There may even be a coaxial cable at the input to the preamplifier, but one is not included in this discussion. Adding an input cable doesn't change the results of the analysis, since as we shall see, the interfacing component characteristics do not enter into the measurement calculations.

The components do have to be selected to work well at the measurement frequency, though. In addition, the preamplifier must be chosen so that with its input terminated, the output noise power is higher than the spectrum analyzer's noise floor. The measurement circuit can be represented as a black box with gain (G_M), effective input noise temperature (T_M), and the display.



2. These measurement tests are required to eliminate test equipment effects on the DUT noise power measurement.

ERADICATE MEASUREMENT CIRCUIT EFFECTS

Eliminating the test circuit effects on the measured noise power from the DUT requires that we make two additional measurements without the DUT. *Figure 2* shows the required measurements. They consist of two noise reference sources measured at the input to the test measurement circuit. The noise sources can be two separate sources, or one source with the second source made up of the first source with an accurately known attenuator on its output.

The black box representation shown in *Figure 1* allows us to simplify the analysis. Furthermore, we can write much more efficient equations that enable the determination of the actual noise power present at the input of the test circuit.

If we terminate the input to the test circuit with two noise sources of different excess noise ratio (ENR), then the two corresponding equations for the displayed noise power are:

$$\text{Displayed noise power on the spectrum analyzer} = (T_{N1} + T_M) \times G_M \text{ and } (T_{N2} + T_M) \times G_M \quad (1)$$

In the equations, T_{N1} and T_{N2} are the two noise-source temperatures. G_M and T_M are the equivalent measurement circuit gain and noise temperature parameters, respectively.

We now form two Y-Factors using the powers given in Equation 1 with the output power from the DUT, which we will denote as T_{UNK}.

Note: A Y-Factor is simply the ratio of two power measurements as read on the spectrum analyzer. It can be stated as either a numerical ratio, or as dB. The context will dictate which one is used.

The two Y-Factor equations are:

$$Y_1 = \frac{(T_{N1} + T_M) \times G_M}{(T_{UNK} + T_M) \times G_M} \quad (2)$$

$$Y_2 = \frac{(T_{N2} + T_M) \times G_M}{(T_{UNK} + T_M) \times G_M} \quad (3)$$

Here, we have shown two Y-Factors, one being the ratio of the measured output power with the first noise source (we'll call it noise source #1) to the output power with the unknown noise power source. The other is the ratio of the measured output power with the second noise source (noise source #2) to the output power with the unknown noise power source.

We rearrange equations 2 and 3 as such:

$$(Y_1 \times T_{UNK}) + (Y_1 \times T_M) = T_{N1} + T_M \quad (4)$$

$$(Y_2 \times T_{UNK}) + (Y_2 \times T_M) = T_{N2} + T_M \quad (5)$$

Rearranging:

$$T_M \times (Y_1 - 1) = T_{N1} - (Y_1 \times T_{UNK}) \quad (6)$$

$$T_M \times (Y_2 - 1) = T_{N2} - (Y_2 \times T_{UNK}) \quad (7)$$

Let $k_1 = Y_1 - 1$ and $k_2 = Y_2 - 1$. We also let $m_1 = k_1 \times Y_2$ and $m_2 = k_2 \times Y_1$.

If one uses the subtraction method and makes the above substitutions (for simplification), the result is:

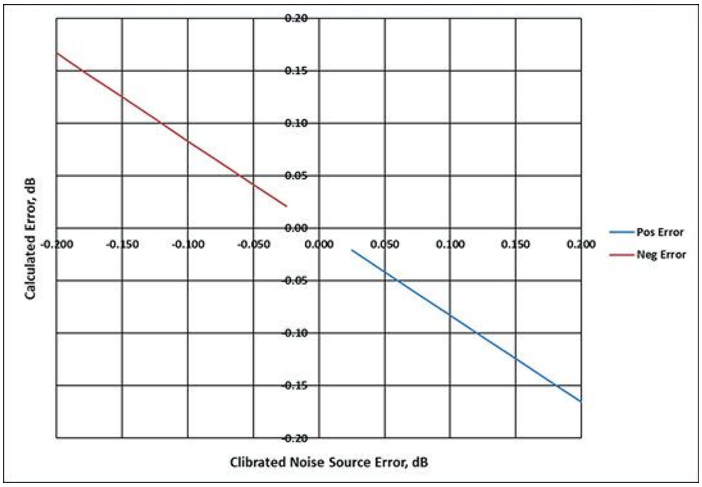
$$T_{UNK} = \frac{(k_1 \times T_{N2}) - (k_2 \times T_{N1})}{m_1 - m_2} \quad (8)$$

All variables in this equation are either measurement quantities, or noise temperature values (related to the noise power by P = kTB) given by the manufacturer of the calibrated noise source. From this equation, we can calculate the noise temperature out of the DUT without knowing the gains and noise figures of the components in the test-measurement circuit.

If we use the thermal model to examine the error associated with the error of the calibrated noise sources, we see that the overall error is essentially the same as the noise sources *(Fig. 3)*.

APPLYING TO REAL-WORLD PROBLEMS

Now that we have a measurement tool that can measure the precise power level of an unknown noise source (ignoring the error associated with the calibrated noise source), we can apply it to the two original problems.



3. Shown is the measurement error associated with noise source error.

NOISE-FIGURE MEASUREMENT

Noise Factor (which is related to the noise figure by: NFac = $10^{(NFig/10)}$) is defined by Equation 9, which states that the noise factor of a component is equal to the Enr (Enr = $10^{(ENR/10)}$, where ENR is in dB) of the calibrated noise source divided by the measured quantity "Y-Factor-1." In this context (classical noise-figure measurement method), Y-Factor is the ratio of the DUT's output power with a calibrated noise source connected to its input (this is called the hot noise source) to the DUT's output power when its input is terminated with a matched load (this is called the cold noise source).

$$\text{Noise Factor} = \frac{\text{Enr}}{(Y - 1)} \qquad (9)$$

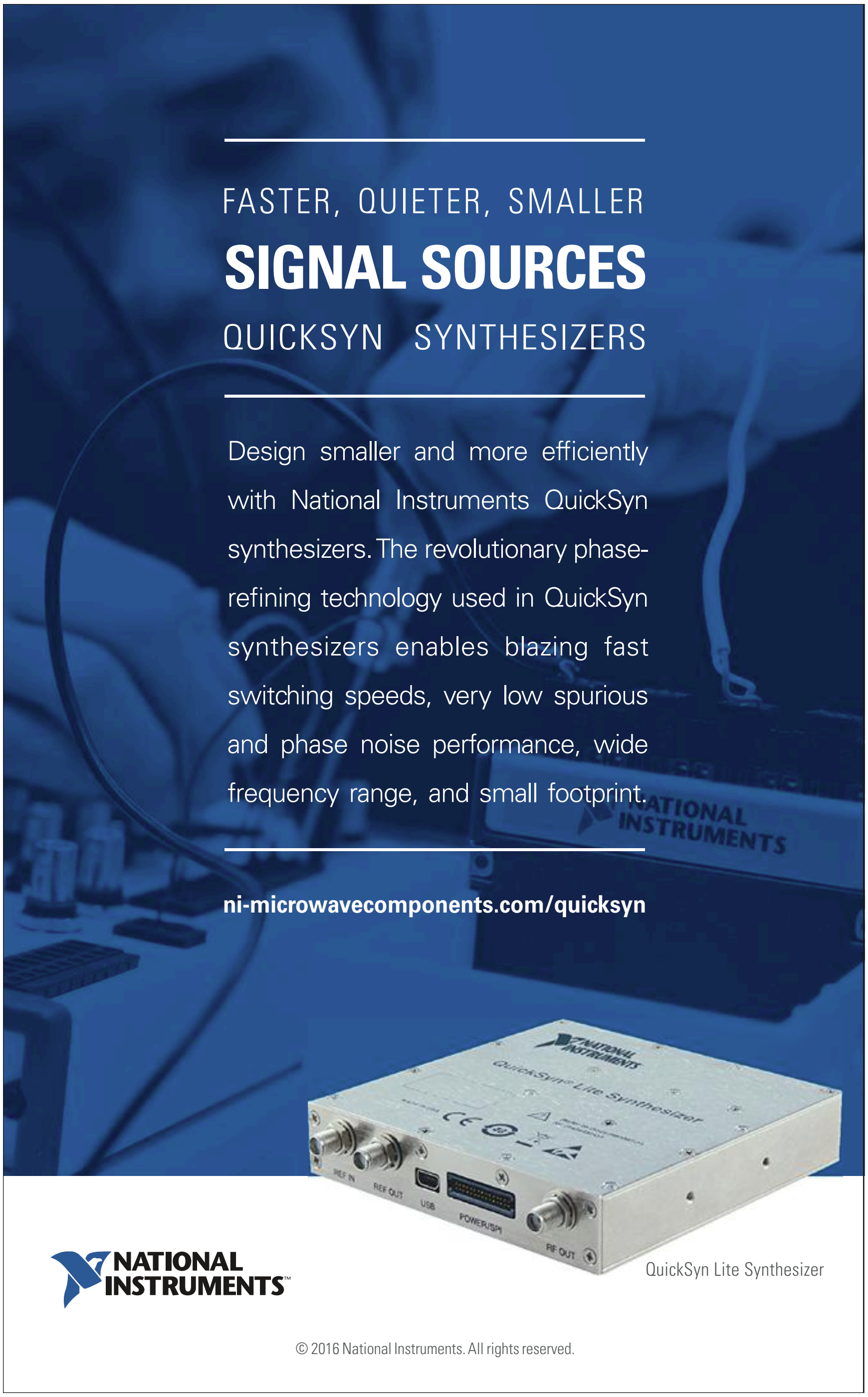
This formula is valid when the noise out of the DUT significantly dominates all of the noise contributions of the test-measurement circuit. This will hardly be the case, which leads to the many accuracy issues that can arise. If we use the noise-power measurement technique discussed in this article, we can measure the output power of the DUT for the hot/cold conditions and precisely determine the ratio at the output of the DUT.

The two unknown temperatures, represented by the T_{UNK} in Equation 8, are:

- Output from the DUT with a cold noise termination = T_{UC}
- Output from the DUT with a hot noise termination = T_{UH}


PHA
PMA
LHA
PHA
PMA
PHA

> It's worth mentioning that the attenuator used to create a lower Enr noise source needs to be accurately measured. Any error in this measurement will propagate to the results.



Now that we have measured the appropriate Y-Factors to calculate the output noise powers out of the DUT, we can calculate the noise factor of the DUT from the classic formula given in Equation 8:

$$\text{Noise Factor} = \frac{\text{Enr}}{(T_{UH}/T_{UC} - 1)} \qquad (10)$$

From which the noise figure follows:

$$\text{NFig} = 10 \times \text{Log}_{10}(\text{NFac}) \qquad (11)$$

NOISE-FIGURE MEASUREMENT SUMMARY

To summarize the process, we need to make four noise power measurements read on the spectrum analyzer:

1. A calibrated noise source connected directly to the test measurement circuit input. This would be connected to the buffer amplifier shown in *Figure 2*. Call this P1.
2. A second calibrated noise source (or the first source with a high precision, measured attenuator at its output) connected to the buffer amplifier shown in *Figure 2*. Call this P2.
3. The DUT with its input terminated in a matched load (290K) connected to the test circuit. Call this P3.
4. The DUT with its input terminated with a calibrated noise source connected to the test circuit. Call this P4.

Next, to calculate the power coming out of the DUT, we form four Y-Factors:

1. $Y_1 = P_1/P_3$
2. $Y_2 = P_2/P_3$
3. $Y_3 = P_1/P_4$
4. $Y_4 = P_2/P_4$

Using Equation 8 and Y_1, Y_2, T_{N1}, and T_{N2} to calculate k_1, k_2, m_1, and m_2, we calculate the temperature out of the DUT for the cold noise case. Call this T_1.

Using Equation 8 and Y_3, Y_4, T_{N1}, and T_{N2} to calculate k_1, k_2, m_1, and m_2, we calculate the temperature out of the DUT for the hot noise case. Call this T_2.

Form the Y-Factor, $Y_5 = T_2/T_1$.

Using the classical Equation 9, we obtain the result:

$$\text{Noise Factor} = \frac{\text{Enr}}{(Y_5 - 1)} \qquad (12)$$

where Enr is the excess noise ratio of the calibrated noise source used for the hot source into the DUT.

The entire process can be accomplished by using a single calibrated noise source. It's worth mentioning that the attenuator used to create a lower Enr noise source needs to be accurately measured. Any error in this measurement will propagate to the results. Fortunately, high-accuracy attenuation measurements can easily be made with today's test equipment.

One last important note: While it's true that the exact component characteristics of the interfacing amplifier, cables, and spectrum analyzer are not used in the calculations for the DUT noise figure, one still must be careful choosing the components of the measurement circuit. The measurement circuit must be constructed such that a 290 Kelvin termination at the input to the DUT is above the spectrum analyzer's noise floor relative to when the test circuit is terminated in a matched load (290 Kelvin). There's no restriction on how far above the test circuit noise floor it must be—it just must be measurable.

This is the issue addressed by this article, which presents a method that eliminates the effects of the analyzer and its connecting circuitry. The result is a noise-figure measurement of the DUT irrespective of the circuitry on its output.

HIGH-POWER NOISE SOURCE

Applying the measurement technique to calibrate a high-power noise source should be clear at this time from the above discussion, but it will be briefly explained.

Many devices can generate large levels of noise power. An amplifier will produce "290 × Gain" white noise at its output over the bandwidth it was designed for. Many other devices also can produce quite a bit of noise (we're usually trying to reduce the noise). If you want a very high-power noise source, you can cascade amplifiers to greatly increase the noise power output.

It should be noted that when driving amplifiers, receivers, etc., from a wideband high-power noise source, care must be taken not to saturate the receiving test circuit. An output filter can be used to band-limit the output and keep the total integrated noise power to a level that will not saturate whatever is being driven from the noise source.

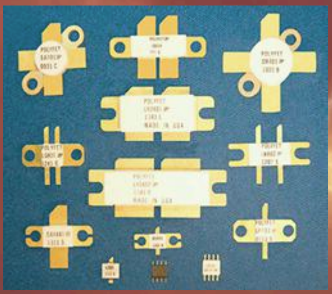
Once the noise source is obtained, its output can be accurately measured using the technique discussed in this article to provide a noise source that's as accurate as the calibrated noise source used for reference.

REFERENCES

1. "Fundamentals of RF and Microwave Noise Figure Measurements," Application Note, Keysight Technologies.

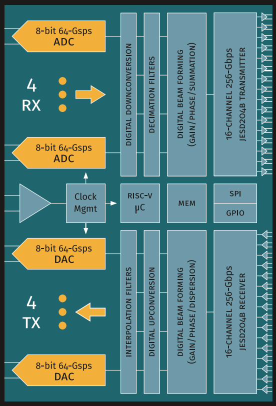
2. "Noise Source Components," Mercury Systems Tech Brief.

> While it's true that the exact component characteristics of the interfacing amplifier, cables, and spectrum analyzer are not used in the calculations for the DUT noise figure, one still must be careful choosing the components of the measurement circuit.





BEYOND WHAT'S NEXT ™
IQ-ANALOG
SCAR
SGT WEBER
PV2 REID
2 CAV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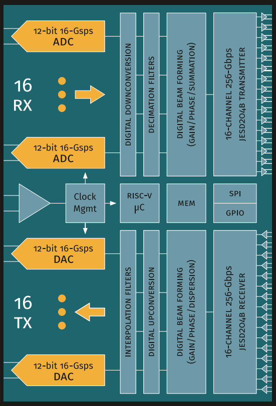
4 RX
8-bit 64-Gsps ADC
DIGITAL DOWNCONVERSION
DECIMATION FILTERS
DIGITAL BEAM FORMING (GAIN/PHASE/SUMMATION)
16-CHANNEL 256-Gbps JESD204B TRANSMITTER
Clock Mgmt
RISC-V μC
MEM
SPI
GPIO
8-bit 64-Gsps DAC
4 TX
8-bit 64-Gsps DAC
INTERPOLATION FILTERS
DIGITAL UPCONVERSION
DIGITAL BEAM FORMING (GAIN/PHASE/DISPERSION)
16-CHANNEL 256-Gbps JESD204B RECEIVER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defense electronics

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BAE BOOSTS F-35's
EW Capabilities

JACK BROWNE | Technical Contributor

ALREADY THE WORLD'S most advanced fighter aircraft, the fifth-generation F-35 Lightning II is about to receive new capabilities for its on-board electronic equipment. The enhancements come courtesy of a Block 4 Modernization contract award from Lockheed Martin *(www.lockheedmartin.com)* to BAE Systems *(www.baesystems.com)*. Under the contract, BAE will enhance the offensive and defensive capabilities of the F-35's AN/ASQ-239 Electronic-Warfare/Countermeasures (EW/CM) systems to address emerging threats.

"The F-35 will be in service for decades, and we're committed to providing our pilots with an AN/ASQ-239 capability that affords a decisive and sustained EW operational advantage," said Deborah Norton, vice president of F-35 Solutions at BAE Systems. "Our robust, modular architecture enables us to efficiently insert new capabilities, supporting the next wave of technical innovation while proactively addressing total product lifecycle sustainability."

BAE Systems has been the source for EW systems for the F-35 *(see figure)* for the past 14 years—it's designed and developed the Block 1, Block 2, and Block 3 configurations. The Block 4 program is a multiyear, multiple-contract design-and-development effort to add 11 new capabilities to F-35 Lightning EW suites.

The enhancements will be made as part of continuous capability development and delivery (C2D2) efforts, in which capabilities are developed and integrated in six-to-twelve-month intervals to keep pace with evolving threats. BAE has delivered more than 500 F-35 AN/ASQ-239 EW/CM systems, bringing more than 60 years of EW technology development and meeting on-time delivery demands with modern production methods and technologies. de

The fifth-generation F-35 Lightning II is about to receive new EW capabilities from BAE Systems.

(Courtesy of Lockheed Martin)

Lockheed Martin Protects Navy from Missile Attacks

DETECTION AND RESPONSE are key functions in any missile defense system, especially for aircraft carriers at sea. The U.S. Navy did not make the choice of its carrier defensive system lightly, selecting Lockheed Martin *(www.lockheedmartin.com)* for the next Ship Self-Defense System (SSDS) Combat System Engineering Agent (CSEA). The SSDS is a combat detection system that protects against anti-ship missile attacks for aircraft carriers and amphibious large-deck ship classes. As part of the contract, Lockheed Martin will evolve the SSDS combat system, beginning with SSDS Build 12, making new capability upgrades and maintaining SSDS in-service baselines.

"The Lockheed Martin CSEA team recognizes the critical role the Aircraft Carrier plays in American power projection and the central role Large Deck Amphibious Ships serve in defense of our nation and our allies," said Jim Sheridan, vice-president, Naval Combat and Missile Defense Systems, Lockheed Martin. "These missions require the highest standards of

(Continued on page 48)

In Search of Accurate Robotic Decision-Making

JACK BROWNE | Technical Contributor

MUCH HAS BEEN MADE in recent years how artificial intelligence (AI) is changing our ways of life in all environments, from commercial practices to electronic warfare (EW). Seemingly more than one-half of today's telephone calls are being made by automated, robotic marketing systems, while almost every new application for credit of any kind questions whether an applicant is a robot.

Robots are performing more of the repetitive tasks in day-to-day life that humans once considered boring, such as making those repetitive marketing phone calls. And, in some cases, such as calls for worthwhile needy charities, those robotic calls can bring a world of good to humans.

Robotic control of autonomous "self-driving" vehicles, such as those guided by advanced driver assistance systems (ADAS), is being projected to make future highways safer as more and more drivers take their hands "off the steering wheel." Some vehicle manufacturers have even presented plans for autonomous vehicles without steering wheels. They expect that future self-driving vehicles, guided by on-board computers with an abundance of navigational data from vehicular radars and other sensors, will treat drivers more as passengers, resulting in much safer highways.

But what about on the battlefield? What about when faced with complex situational awareness issues and complicated operating environments, when robots and humans team up as troops to face an adversary that also presents armed forces that are a mix of human and robotic soldiers?

Classic science-fiction television shows such as the late 1960s *Star Trek* series envisioned warfare with man and machine closely interacting, often relying on the starship's central computer for guidance. The two central characters, very human and emotionally charged Captain James T. Kirk and logic-driven, half-human/half-Vulcan co-pilot Mr. Spock, often debated critical decisions with the ship's computer as mediator. As is proving true, modern AI technology is helping to develop robotic partners that bring a high degree of logic to the battlefield.

Many nations, including the U.S. and China, are actively pursuing major roles for robotic systems on the battlefield.

(Continued on page 66)

Lockheed Martin Protects Navy from Missile Attacks

(Continued from page 45)

systems engineering, performance analysis, software development, integration, and test to deliver an integrated Ship Self-Defense System. We are proud to have been selected by the Navy for this important program."

The fifth-generation F-35 Lightning II is about to receive new EW capabilities from BAE Systems. *(Courtesy of Lockheed Martin)*

Lockheed Martin has considerable experience in providing integrated combat systems for U.S and international navies, serving as the Aegis CSEA. The company developed the Common Source Library (CSL), which enables efficient deployment of common software solutions across the Surface Navy, with variation techniques to customize different configurations. The approach has made it possible to deliver weapons systems focused on the needs of the U.S. Coast Guard and Littoral Combat Ship that were derived from the technology employed in the U.S. Navy's Aegis systems. "We are excited to extend our Aegis CSEA and Frigate Weapon System experience to become the SSDS CSEA," added Sheridan. de

Rapid Prototyping Enhances Tactical Vehicles

DESIGN KITS ARE HELPING to streamline the installation of advanced logistics and secure communications systems into tactical vehicles in the field. The U.S. Army Futures Command (AFC) and its Joint Battle Command-Platform (JBC-P) are speeding systems installations via 14 different design kits used for the Family of Medium Tactical Vehicles (FMTVs). The AFC has been developing universal design kits that can be applied across many different tactical vehicles and across joint forces.

""The process has been iterative with designing, testing, and validating. We've been prototyping as we go," says Tim Knabel, a project lead in the Prototyping Integration Facility of CCDC's center for Command, Control, Communications, Computers, Cyber, Intelligence, Surveillance, and Reconnaissance (C5ISR).

The kits were designed for high-volume manufacturing processes, so that thousands of kits could be produced and used across many different types of vehicles. The prototypes were designed to meet all technical specifications by testing for mechanical, electrical, temperature, humidity, shock, and vibration requirements. Knabel also noted the value of feedback from soldiers *(see figure):* "We're focused on understanding the needs of soldiers; this incorporates their input throughout the project, which leads to a better solution and improved soldier functionality."

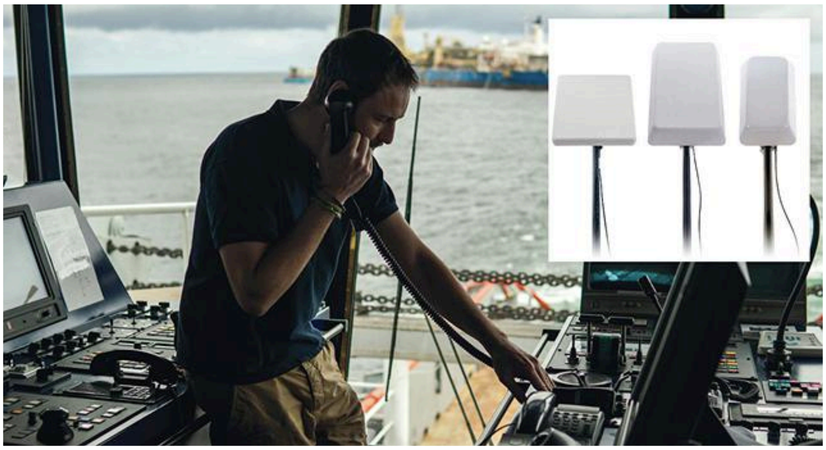
Human systems integration was also at the forefront of the process, according to Mark Krivansky, a Product Manager JBC-P industrial engineer. "The design has taken into consideration human factors and ergonomics with the goal of reducing human error, increasing the soldier's productivity, and enhancing safety and comfort," says Krivansky. "Soldiers have been incorporated as one of the elements of the design, with a focus on how they interact successfully with the JBC-P system. Now, the entire system will be mounted directly in front of soldiers, for full access and usage, while not interfering with visual assessment of their environment, both inside and outside the vehicle." ■

The Army Futures Command is developing design kits that can help quickly upgrade the electronic capabilities of tactical vehicles. *(Courtesy of the U.S. Army Research Laboratory)*


CELEBRATING
50
YEARS
UNBEATABLE
QUALITY
SERVICE
STABILITY &
INNOVATION

MANET Radio Antennas Install in Minutes

MILITARY AND CIVILIAN USERS of Wave Relay mobile ad hoc network (MANET) radios now have an easy-to-install companion antenna system from Persistent Systems LLC *(www.persistentsystems.com)*—the Integrated Antenna Series. Available in S- and C-band sector antennas as well as C-band directional antennas, these reliable antenna systems *(see figure)* install quickly (within minutes in most locations) using just basic tools. The directional antenna links communication towers to form wireless backbone networks. The sector antennas provide the long range and high throughput rates needed for large coverage areas. Data routing is handled by the Wave Relay MANET radio integrated into each antenna.



These S- and C-band antennas integrate radios and install quickly, thus creating radio links in minutes.

"These tower antennas can support a variety of MPU5 MANET radio users, such as military convoys, wildland firefighters, test range personnel, law enforcement, port security—and especially border security," says Nick Naioti, vice president of Business Development at Persistent Systems. "Applications like border security require users to operate across thousands of square miles and remain continuously connected to each other and the enterprise. That means continuous access to live camera feeds and sensor data to positively impact the mission."

Designed for installation in just a few minutes, the new Integrated Antenna Series has only one Ethernet cable that provides both power and data connectivity. In addition, the MANET radio is housed in an IP66-rated enclosure on the back of the antenna. ■

MEMS Technology Fuels Timing Solutions

A LINE OF PRECISION TIMING SOLUTIONS based on microelectromechanical-systems (MEMS) technology have been developed to handle harsh conditions in defense and aerospace systems. The frequency-reference sources—the Endura timing solutions from SiTime Corp. *(www.sitime.com)*—can be specified for a number of different performance parameters, including operating frequency, frequency and phase stability, and power consumption, as needed for in-field and space systems with long-term timing requirements *(see figure)*.

"When exposed to high levels of shock, vibration, and extreme temperatures, legacy timing components have been prone to failure, degrading system performance and reliability," says Piyush Sevalia, executive vice president of marketing. "To solve these problems, SiTime created an oscillator system of silicon MEMS, analog circuits, compensation algorithms, and advanced packaging, which is designed to outperform any other available timing solution in harsh environments. For example, Endura precision TCXOs deliver 4 parts per trillion per g (ppt/g) of acceleration sensitivity, which is 50 times better than legacy quartz-based solutions. With such performance, we believe that Endura will transform the oscillator landscape in aerospace and defense."

As an example, the Endura SiT8944 oscillator has an available frequency range of 1 to 110 MHz with low acceleration sensitivity of 0.1 ppb/g and an operating temperature range of −55 to +125°C. It's designed for shock levels to 10,000 g's and provides 70 g vibration resistance. The device fits within a compact 2.0- × 1.6-mm package and conforms to MIL-PRF-55310, MIL-STD-883, and AEC-Q100 specifications. ■

Based on MEMS technology, these frequency-reference sources can endure harsh shock and vibration conditions for reliable long-term military and aerospace systems.

NEL FREQUENCY CONTROLS, INC.

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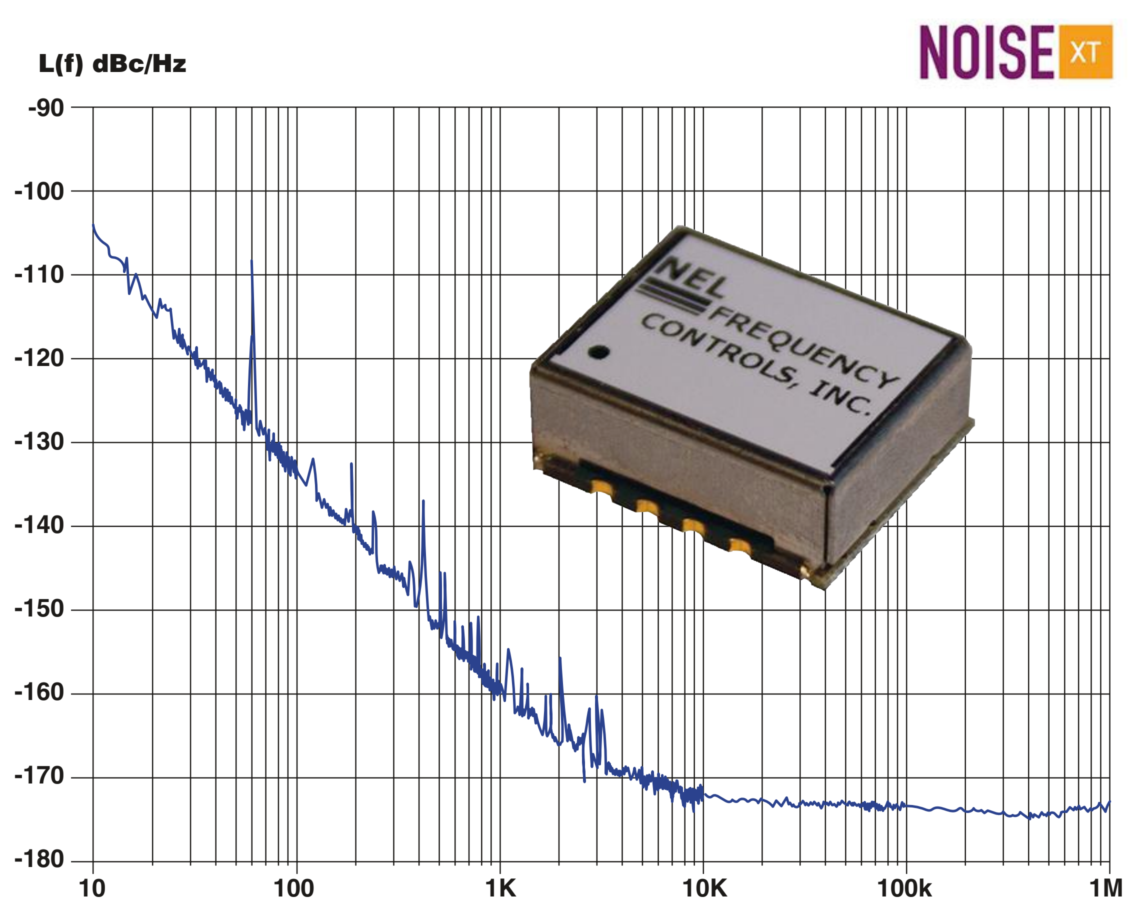
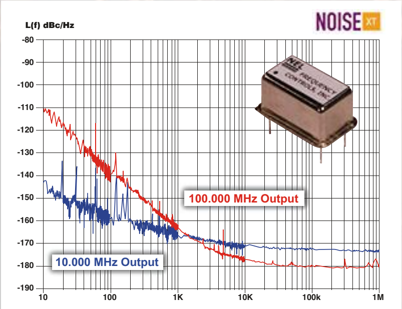
Ultra Low Phase Noise Frequency Control Products

ULPN TCXO @ 100 MHz with Low G Sensitivity 0.2ppb/ G

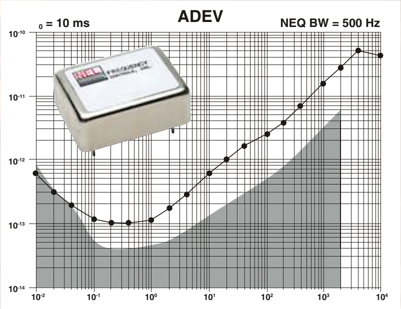


DIP 14 OCXO— 10 MHz or 100 MHz



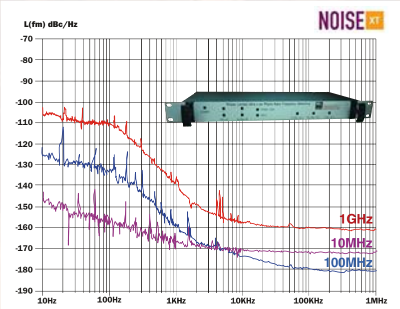
Ultra low phase noise, low power consumption (250 – 350mW)

ULPN OCXO @ 10 MHz



Outstanding close to carrier phase noise and short term stability

ULPN 10/100/1000 MHz Appliance



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JACK BROWNE | Technical Contributor

MODULAR SYNTHS That Meet Today's Tough Specs

Trends in smaller and lighter frequency synthesizers for defense and aerospace applications translate to fitting more performance at higher frequencies into modular formats.

FREQUENCY SYNTHESIZERS provide the RF/microwave signals so essential to the operation of many electronic defense systems. They have steadily evolved over time to provide wider bandwidths through the millimeter-wave (mmWave) frequency range, meeting growing demands for smaller step sizes (frequency resolution) with lower phase noise and faster frequency switching speeds. Recent trends for miniaturization and modularity have increased the flexibility when integrating a frequency synthesizer into a system, especially where size, weight, and power (SWaP) are concerns for smaller systems like portable radios and unmanned aerial vehicles (UAVs).

1. Fully equipped modern wideband frequency synthesizers can fit within a single PXI slot. ***(Courtesy of Micro Lambda Wireless)***

The process of specifying a synthesizer has not changed—it's still driven by the requirements of an application and comparing key performance parameters, such as frequency range, tuning resolution, frequency switching speed, phase noise, harmonics, and SWaP. Whether in a 19-in. rack-mount assembly or a more compact module, frequency synthesizers for military and aerospace applications must perform when needed, no matter how severe the temperature, shock, and vibration conditions.

Specifying a synthesizer usually starts with frequency range. High-frequency synthesizers are available with a single fixed frequency on up to extremely wideband models covering several octaves. Choosing the right source for an application usually involves running a checklist of performance parameters. Some of those parameters may not be a consideration, such as fast switching speed for a single-frequency synthesizer. However, combinations of other performance parameters, such as low phase noise, harmonics, and spurious levels, may be critical for the same synthesizer.

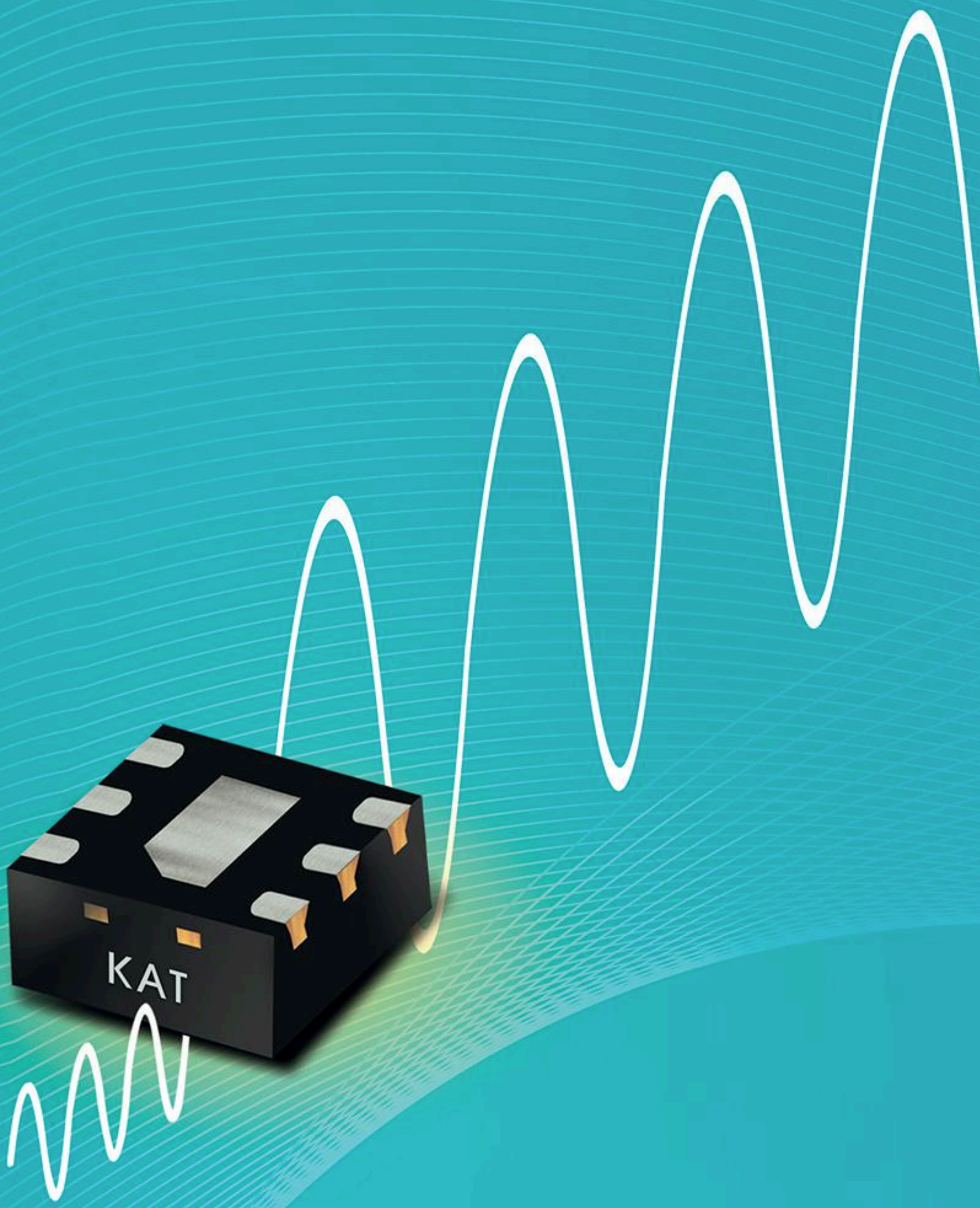
MOVING TO MODULAR

Increasingly, portable/mobile defense/aerospace applications in communications, surveillance, and guidance systems are driving the development of smaller and lighter frequency synthesizers. It's thus boosting the interest in smaller frequency-synthesizer modules for many applications, including in test-and-measurement systems.

Frequency synthesizers can be designed in many mechanical formats, from tiny surface-mount-technology (SMT) packages to legacy rack-mount enclosures for test-and-measurement and other system applications. Modular formats, such as standard VME and VPX modules with multiple interfaces for digital control, have become popular for the flexibility they bring to system installation.

Frequency synthesizers are constructed in several configurations. These include indirect formats in which low-noise output signals are generated through the combination of a reference oscillator with a higher-frequency tunable oscillator, such a dielectric resonator oscillator (DRO), voltage-controlled oscillator (VCO), or yttrium-indium-garnet (YIG) oscillator.

Using a loop filter, phase comparator, and phase-locked loop (PLL) to com-


KAT

pare the phase of the two oscillators, the frequency accuracy and phase stability of the reference source can be translated to the RF/microwave oscillator. Fractional-N or integer-N frequency division/multiplication provides frequency resolution as a function of the reference oscillator. The reference oscillator is usually a crystal oscillator, although atomic oscillators such as rubidium (Ru) or cesium (Cs) standards have also served as references.

Direct digital synthesizers (DDSs) take advantage of ever-faster digital circuitry and digital components such as field-programmable gate arrays (FPGAs) and digital-to-analog converters (DACs), to directly generate RF/microwave waveforms from digital codes. The spectral purity of a DDS is largely determined by the DAC's performance, such as its bit resolution, although the phase noise is a function of the reference oscillator. DDS techniques are particularly effective when multiple phase-matched or related output signals must be generated, e.g., in-phase (I) and quadrature (Q) signals for differential receivers.

SPECIFYING SOURCES

The "move to modular" in frequency synthesizer design is apparent with some of the latest product designs from leading suppliers, such as Fairview Microwave, FEI-Elcom *(see page 66)*, Mercury Systems, Micro Lambda Wireless, Pasternack, Phase Matrix *(www.phasematrix.net*, a wholly owned subsidiary of National Instruments, *www.ni.com)*, Teledyne Microwave Solutions *(www.teledynemicrowave.com)*, and Textron Systems. Even smaller synthesizers, including those in surface-mount-technology (SMT) housings, are available from a number of companies, including Analog Devices *(www.analog.com)* and Synergy Microwave Corp. *(www.synergymwave.com)*.

For example, Micro Lambda Wireless *(www.microlambdawireless.com)*, a long-time producer of YIG-based PLL frequency synthesizers, squeezes wideband frequency synthesizers through mmWave frequencies into modules occupying a single PXI chassis slot *(Fig. 1, page 52)*. The firm's MLMS-Series of wideband frequency synthesizers covers standard frequency ranges of 0.25 to 32 GHz with tuning resolution of 1 kHz when operating with an internal crystal reference oscillator.

Although YIG-based PLL frequency synthesizers sacrifice tuning speed (compared to VCO-based frequency synthesizers), with frequency tuning speeds of 1 to 3 ms, they are noteworthy for their low phase noise, as needed for sensitive receivers. The MLMS-Series PXI modular synthesizers provide just that, with very low phase noise through 32 GHz.

2. These compact modular frequency synthesizers are optimized for EW and ELINT systems. *(Courtesy of Mercury Systems)*

For carriers from 250 MHz to 6 GHz, the single-sideband (SSB) phase noise is −75 dBc/Hz offset 100 Hz from the carrier, −94 dBc/Hz offset 1 kHz, −119 dBc/Hz offset 100 kHz, and −142 dBc/Hz offset 1 MHz from the carrier. For carriers from 28 to 32 GHz, the SSB phase noise is −63 dBc/Hz offset 100 Hz from the carrier, −83 dBc/Hz offset 1 kHz, −113 dBc/Hz offset 100 kHz, and −137 dBc/Hz offset 1 MHz from the carrier. These performance levels are measured for synthesizers operating with an internal crystal reference oscillator. The synthesizers can also be supplied with external frequency reference sources, with options for 50 to 200 MHz or 10 to 200 MHz, both at nominal input power of 0 dBm.

Standard units are designed for operating temperatures from 0 to +60°C, although custom units are also available for a wider operating temperature range of −40 to +85°C. Measuring just 2.5 × 2.5 × 0.65 in., they deliver healthy output levels from +8 to +13 dBm. Spurious levels range from −54 to -60 dBc, depending on frequency, while second harmonics are typically −8 to −20 dBc.

When frequency switching speed is more of an issue, the firm also provides VCO-based frequency synthesizers as part of its MLVS-Series LUXYN frequency synthesizer, covering a total frequency range of 50 MHz to 21 GHz. The VXI modular synthesizers achieve full-band tuning speed of 50 μs with as much as +15 dBm output power. They exhibit −125 dBc/Hz SSB phase noise offset 10 kHz from a 10-GHz carrier.

DS-3000 frequency synthesizer modules from Mercury Systems *(www.mrcy.com)* are optimized for EW and ELINT applications from 100 MHz to 20 GHz *(Fig. 2)*. The DDS-based frequency synthesizers use list-mode tuning under four-wire SPI control for tuning speeds to 200 μs with 1-Hz frequency tuning resolution. They include mini-USB connectors and female SMA RF connectors. The synthesizer modules measure just 6.5 × 4.0 × 0.7 in. and weigh 14.5 oz. with maximum power consumption of 12 W (typically 10 W) for an operating temperature range of −30 to +70°C. They exhibit −12 dBc harmonic levels with −60 dBc maximum spurious levels.

The two versions of synthesizer modules demonstrate the impact of reference frequency stability on synthesizer performance. The DS-3001 is equipped with an internal 100-MHz temperature-compensated crystal oscillator (TCXO) reference with better than 2 ppm aging rate and ±0.5 ppm stability with temperature. The phase noise is typically −76 dBc/Hz offset 100 Hz from a 10-GHz carrier, −106 dBc/Hz offset 1 kHz from the same carrier, −121 dBc/Hz offset 10 kHz from

10 GHz, −124 dBc/Hz offset 100 kHz from 10 GHz, and −122 dBc/Hz offset 1 MHz from the same carrier.

When better phase-noise performance is required, the DS-3002 frequency synthesizer is supplied in the same modular format, but with a higher-stability internal 100-MHz oven-controlled crystal oscillator (OCXO) with better than 1-ppm aging rate and ±0.1-ppm stability with temperature. The phase noise with the OCXO reference is typically −96 dBc/Hz offset 100 Hz from a 10-GHz carrier, −114 dBc/Hz offset 1 kHz from the same carrier, −122 dBc/Hz offset 10 kHz from 10 GHz, −124 dBc/Hz offset 100 kHz from 10 GHz, and −122 dBc/Hz offset 1 MHz from the same carrier. Both synthesizer configurations deliver +13 dBm output power with ±2 dB flatness.

3. Both DDS-based and indirect PLL-based frequency synthesizers can be supplied in modular formats. *(Courtesy of Kratos General Microwave)*

When precise frequency resolution and power control are needed in a slightly larger package, Textron Systems *(www.textronsystems.com)* offers a series of frequency synthesizers for EW and automated-test-equipment (ATE) applications in two-slot VXI C size modules. They measure 10.4 × 2.4 × 14.5 in. and weigh 11.5 lbs, controlled by a VXI 3.0 interface. The compact synthesizer modules are available over a total frequency range of 3 MHz to 40 GHz with 0.04-Hz frequency resolution and an output power range of −100 to +18.5 dBm for signals to 20 GHz. Spurious content is −55 dBc at +10 dBm output power. The synthesizers are designed for use with 10-MHz reference oscillators, including external atomic Ru oscillators.

These VXI synthesizers can achieve better than 500-ns switching speed and a wide range of modulation using internal or external sources, including AM, FM, pulse modulation, differential I/Q modulation, binary phase-shift-keying (BPSK) modulation, quadrature phase-shift-keying (QPSK) modulation, and quadrature amplitude modulation (QAM). They are equipped with 2.4-

mm female RF output connectors and female SMA connectors for the external reference oscillator.

Kratos General Microwave *(www.kratosmed.com)* offers direct and indirect frequency synthesizers in modular form *(Fig. 3)*. While the DDS-based frequency synthesizers are capable of extremely fast frequency switching speeds (to 40 ns) with frequency- and phase-coherent switching, the indirect PLL frequency synthesizers can still deliver frequency switching speed as fast as 1 μs between two frequencies. The DDS-based synthesizers are available for tuning from 10 MHz to 18 GHz in 10-MHz steps. For slower switching speeds, the indirect PLL synthesizers provide much smaller frequency step sizes across the same frequency range. The firm offers custom synthesizer designs through 40 GHz.

Some suppliers prefer to work directly with customers to fulfill system requirements for frequency synthesizers rather than offering standard models. Long-time frequency-synthesizer supplier Wide Band Systems *(www.widebandsystems.com)* has a strong track record of providing synthesizers for tactical signal identification and analysis systems around the world. Switching speeds are 0.3 μs and better and spectral purity ranges from 0.5 through 18.0 GHz and beyond for the firm's synthesizers, which come in various mechanical formats including compact modules *(Fig. 4)*.

4. Broad frequency ranges with custom performance requirements are able to be packed into compact modules. ***(Courtesy of Wide Band Systems)***

The low power consumption of these frequency synthesizers (less than 20 W) makes them attractive for a wide range of commercial and defense applications, including jammers, EW systems, and UAV communications systems. Additional modular synthesizer suppliers offering both standard models and extensive custom capabilities include Mini-Circuits *(www.minicircuits.com)*, Synergy Microwave Corp., and Teledyne Microwave Solutions. **mw**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JACK BROWNE | Technical Contributor

VPX Module Packs Pair of Fast Synthesizers

These compact VPX OSA modules each contain two RF/microwave frequency synthesizers with fast tuning speeds and low phase noise to help miniaturize radar, communications, and EW systems.

SMALLER IS often better for many military/aerospace applications, although miniaturization often comes at a cost. Take RF/microwave frequency synthesizers, for example. Smaller signal sources, such as modules developed for portable use, may lack the output power or noise performance of a larger unit, such as a rack-mount frequency synthesizer intended for test-and-measurement applications.

Fortunately, at least one frequency synthesizer supplier, FEI-Elcom Tech *(www.fei-elecomtech.com)*, has accepted the challenges of delivering large performance in smaller packages. The firm recently announced a series of dual frequency synthesizers designed into standardized VPX Open System Architecture (OSA) modules as well as in ruggedized and customized modules.

More RF/microwave frequency synthesizers are being designed as modules that can be combined with similar component-sized assemblies, such as frequency upconverters and downconverters, local-oscillator (LO) sources, modulators, and even clock timing oscillators, to create compact receivers, transmitters, and transceivers for airborne or portable applications (*see page 50*). Due to the small size of the modules, a multitude of functions can be integrated into one VPX equipment chassis, while the efficient bus control architecture allows for high-speed communications among the modules in a VPX chassis.

PACKING A PAIR

A high degree of integration can be found in FEI-Elcom Tech's most-recent synthesizer line. The firm's VPXMS-2500 dual synthesizer VPX modules (*Fig. 1*) leverage the latest monolithic-microwave-integrated-circuit (MMIC), field-programmable-gate-array (FPGA), and digital-signal-processing (DSP) technologies to create fast-switching (microsecond or faster) dual frequency synthesizers that fit within 6U VPX OSA modules. The performance provided by these compact frequency-synthesizer modules once required 6U spaces in a 19-in. rack.

1. The VPXMS-2500 synthesizers fit two frequency sources within standard VPX modules with modulation bandwidths as wide as 2 GHz. *(Courtesy of FEI-Elcom Tech)*

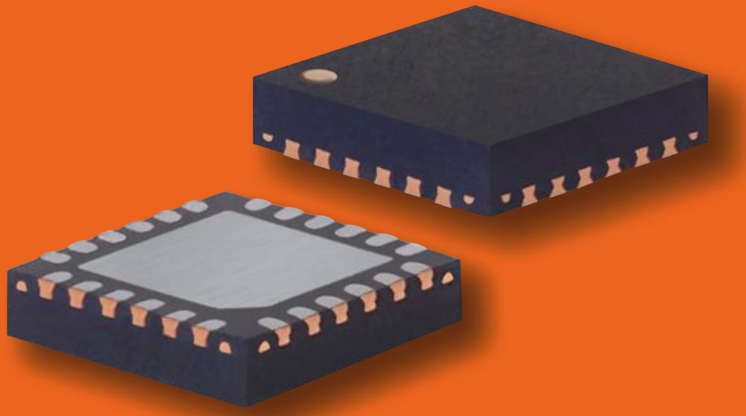
Dual-synthesizer models have been designed in the VPX format for standard frequency ranges of 5 to 10 GHz, 10 to 20 GHz, 5 to 20 GHz, and as wide as 1 to 20 GHz, with other frequency ranges to 67 GHz available as options. The synthesizer modules come with real-time modulation bandwidths as wide as 2 GHz for applications requiring advanced signal generation. They can also be supplied with phase-coherent switching between channels as an option.

Output signals for the dual frequency synthesizer are provided at a robust +15 dBm, flat within ±3 dB. The device features a wideband noise floor of better than −150 dBc/Hz. Its single-sideband (SSB) phase noise is quite low, whether measured close or far from the carrier. Phase noise drops steadily with distance from the carrier. For a 10-GHz carrier, phase noise is −126 dBc/Hz offset 100 kHz from the carrier, −138 dBc/Hz offset 1 MHz from the carrier, −148 dBc/Hz offset 10 MHz from the carrier, and −154 dBc/Hz offset 100 MHz from the carrier.

WHEN ONE IS ENOUGH

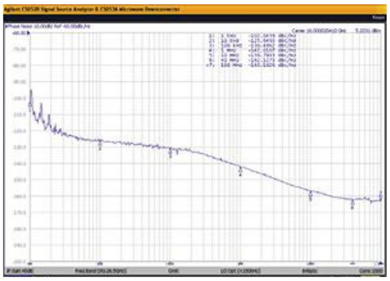
For applications that only need a single frequency source, FEI-Elcom Tech's DFS2000 series of single frequency synthesizers offer outputs in the range from 1 to 20 GHz, also in compact modules (*Fig. 2*). Based on traditional phase-locked-loop (PLL) indirect frequency synthesizer technology and advanced direct-digital-synthesizer (DDS) architectures, these miniature modules provide excellent spectral purity with fast switching speeds from a package measuring just 5.95 × 3.55 × 0.75 in. The DFS2000 series of single-frequency-synthesizer modules are well-suited for test systems and for signal-intelligence (SIGINT) receivers. The modules also provide a fixed low-noise LO signal at L band for receiver applications. The fast-tuning DDS-based synthesizers oper-

2. The DFS2000 series features frequency synthesizers housed in compact modules for space-saving applications from 1 to 20 GHz. *(Courtesy of FEI-Elcom Tech)*



ate from 1 to 20 GHz with a standard tuning step size of 1 MHz (tuning steps as small as 1 Hz and as large as 10 MHz are available as options). High-speed tuning speeds of 10 μs or better can be achieved by using list or trigger sweep modes. The synthesizers excel in terms of noise performance, with harmonics of −20 dBc or better and spurious content of better than −70 dBc within ±100 MHz of the carrier.

Phase noise for the DFS2000 series single synthesizers, as measured with a commercial phase-noise test set from Agilent Technologies (now Keysight Technologies), is quite low, dropping to a noise floor of −150 dBc (*Fig. 3*). For a 20-GHz carrier, the SSB phase noise is −103 dBc/Hz offset 1 kHz from the carrier, −112 dBc/Hz offset 10 kHz from the carrier, −115 dBc/Hz offset 100 kHz from the carrier, −118 dBc/Hz offset 1 MHz from the carrier, and −133 dBc/Hz offset 10 MHz from the carrier.



3. The low phase noise of the DFS2000 series dual frequency synthesizers combines with low harmonics and low spurious noise to deliver excellent spectral purity to 20 GHz. *(Courtesy of FEI-Elcom Tech)*

The frequency synthesizers, which run with typical dc power consumption of 12 W at operating temperatures of −20 to +70°C, are well-suited for radar test, as well as EW and SIGINT receivers. They are designed for use with a reference/clock oscillator at 100 MHz, although synthesizers are available for other reference frequencies as an option.

Frequency synthesizers such as the DFS2000 series sources provide clean signals that can be mixed, multiplied, or divided although they lack the modulation needed for many system applications. The firm offers lines of single frequency synthesizers (without modulation) and signal generators (with modulation) at carrier frequencies from 625 MHz through 67 GHz in a variety of module formats, including standardized modules such as VME and VPX modules. **de**

FEI-ELCOM TECH INC., 260 Union St., Northvale, NJ 07647; (201) 767-8030, FAX: (201) 767-1326, E-mail: sales@fei-elcomtech.com, www.fei-elcomtech.com.

JACK BROWNE | Technical Contributor

When RF/Microwave Connections Are Needed

This long-time connector/adapter source provides the components needed for standard and custom interconnections through 18 GHz and beyond, in coaxial, twinaxial, and triaxial configurations.

HIGH-FREQUENCY RF/MICROWAVE interconnections come in multiple shapes and sizes, and with many different sets of capabilities. They are essential components in most electronic defense and commercial systems and often the last piece in a design puzzle that must process signals for critical battlefield functions, including in radar, electronic-warfare (EW), and electronic-countermeasures (ECM) systems.

RF/microwave connectors that meet required specifications can be difficult to find. However, one source—Connectronics Inc. *(www.connectronicsinc.com)*—has built a solid reputation for over 30 years by supplying high-performance RF/microwave connectors with fast turnaround times and in the amounts needed for the highest-volume projects.

The ISO9001-certified connector developer/supplier provides an extensive line of standard RF/microwave connectors and adapters in addition to many custom-designed connectors. Well-equipped with CNC machinery and RF/microwave test equipment, the company also offers the capabilities to develop high-frequency interfaces based on a customer's mechanical and/or electrical requirements.

Supporting standard signal frequencies mainly from dc through 18 GHz (which are the majority of frequency bands occupied by military/aerospace systems) and with extended frequency coverage through 50.0 GHz for some coaxial connectors and adapters, Connectronics produces both 50- and 75-Ω connectors and adapters in coaxial, twinaxial, and triaxial configurations. Attachment approaches employed include threaded, bayonet, and snap-on (such as in blind-mate connectors).

These are some examples of the wide range of coaxial connectors available in standard and custom configurations. *(Courtesy of Connectronics Inc.)*

STANDARD EXAMPLES

Connectronics produces a wide range of high-frequency connectors. They include lower-frequency UHF and mini-UHF connectors for signal links at frequencies below 1 GHz (*see figure*) to higher-frequency BNC (typically dc to 4 GHz), TNC (typically dc to 11 GHz), and N connectors (typically dc to 11 GHz), as well as SMA and SSMA coaxial connectors with upper frequency limits of 26.5 GHz and beyond. The firm now manufactures 2.4-mm connectors and adapters with low VSWR at frequency limits of 50 GHz.

SMA connectors, probably the most widely used RF/microwave connector in military and aerospace system applications, feature a center conductor surrounded by polytetrafluoroethylene (PTFE) dielectric material, which in turn is surrounded by the outer coaxial metal enclosure. They are typically used for applications from dc to 18 GHz but can support higher-frequency systems when machined to tight-enough tolerances.

SMA connectors are typically tightened by wrench when machined in a knurled-nut configuration. They mate with higher-frequency connectors, such as 2.92- and 3.5-mm coaxial connectors, although they suffer much greater loss than those connectors at their upper-frequency limits (to 40 GHz and beyond). SSMA connectors are smaller versions of SMA connectors, with essentially the same low loss and VSWR through about 18 GHz, for applications with tighter space requirements. Both SMA and SSMA connectors are designed for use with 0.086-in.-diameter semi-rigid cable and can be designed in right-angle and straight (in-line) mounting configurations.

In addition to manufacturing standard RF/microwave connector types, such as for just-in-time (JIT) applications, Connectronics offers an extended line of RF/microwave coaxial adapters plus a healthy assortment of machined components including contacts, housings, insulators, and other special-use parts that are manufactured to customer specifications. With the firm's advanced CNC manufacturing capabilities and test laboratories, special connectors, either based on modifications to standard connector types or designed from a customer's mechanical and electric requirements, can be designed and produced according to the tightest deadlines. de

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CONNECTRONICS INC., 908 S. Walnut St., P. O. Box 246, Edinburgh, IN 46124; (812) 526-8801, E-mail: sales@connectronicsinc.com, www.connectronicsinc.com.

JACK BROWNE | Technical Contributor

Make the Most of Materials at mmWave Frequencies

As frequencies get higher, circuit materials must meet a special set of requirements to optimize designs in the mmWave frequency range.

AS WIRELESS APPLICATIONS continue to expand, uncluttered RF and microwave frequencies are quickly vanishing, inviting circuit designers to look higher. Millimeter-wave (mmWave) frequencies (30 to 300 GHz) offer plenty of available bandwidth if practical components, starting with printed-circuit-board (PCB) materials, can be found to process such high-frequency signals. Fortunately, several materials from Rogers Corp. *(www.rogerscorp.com)* fit the mold for low-loss circuit projects through 77 GHz and higher, including RO3003 and RO4350B circuit materials.

For mmWave circuits, those higher frequencies mean much smaller signal wavelengths than at lower frequencies, requiring PCB materials that can deliver performance goals beyond the levels required at lower frequencies. Insertion loss is one of the key performance parameters at mmWave frequencies, where signal power tends to be much more difficult to generate and maintain than at lower frequencies.

In terms of PCB materials, insertion loss comprises four loss components: conductor loss, dielectric loss, radiation loss, and leakage loss. Conductor and dielectric losses are usually the loss components of main interest when specifying mmWave PCB materials, with the goal of minimizing transmission-line losses though the PCB.

Conductor losses are a function of a PCB material's copper quality, as well as the type of transmission line, such as microstrip or grounded coplanar waveguide (GCPW), used in the circuit design. Dielectric losses are related to the dissipation factor (DF) of a PCB material, with lower DF values translating to lower dielectric loss for a material.

Because of the small wavelengths, thinner PCB materials (with low associated radiation losses) are usually preferred for mmWave circuits. To avoid resonances with mmWave microstrip circuits, it's generally safe to use a PCB material that's thinner than one-eighth of the wavelength of the highest frequency—typically about 5/8 mm at 60 GHz, which has a wavelength of 5 mm. Similarly, the conductor widths of the transmission lines should be less than one-eighth of the wavelength of a circuit's highest operating frequency.

The material thickness and conductor widths are less critical in producing resonances with GCPW transmission lines. However, circuits with GCPW typically suffer higher conductive loss at mmWave frequencies than circuits using microstrip transmission lines.

TIGHT TOLERANCES

Specifying a circuit material for mmWave applications requires close attention to the material's tolerances, such as for conductor and dielectric thicknesses and dielectric constant (Dk). For example, to maintain consistent transmission-line impedance at the small wavelengths of mmWave frequencies, the Dk tolerance should be as tight as possible. Since Dk tolerance also impacts the phase of a high-frequency transmission line, a PCB material with extremely tight Dk tolerance will also support circuits requiring phase matching or minimal phase variations.

A related material parameter, temperature coefficient of dielectric constant (TCDk), provides details on the stability of a circuit material's Dk with changes in temperature. Ideally, a circuit material for military and aerospace mmWave applications will exhibit minimal changes in Dk with environmental changes, such as temperature and relative humidity (RH).

RO3003 and RO4350B circuit materials from Rogers Corp. are examples of circuit laminates with properties well-suited for commercial and military mmWave applications. RO3003 laminate is a PTFE-based circuit material with ceramic filler. It has a Dk of 3.00 at 10 GHz held to a tolerance of ±0.04 across a laminate panel. The Dk is stable with frequency though mmWave frequencies. The DF is low, 0.0010 at 10 GHz, which is a sign of low dielectric loss.

The material's low moisture absorption of 0.04% indicates that the Dk will remain constant even under conditions of high RH. A low TCDk of −3 ppm/°C ensures that the circuit material will maintain a stable Dk value (and transmission lines with stable impedance and phase) even over a wide temperature range of −50 to +150°C.

RO4350B hydrocarbon/ceramic laminate has glass fabric reinforcement. It features the low loss needed for mmWave applications, but it's also RoHS compliant and can be processed in the same way as low-cost FR-4 circuit materials. RO4350B laminate has a Dk of 3.48 in the z-axis (thickness) at 10 GHz, held to a tight tolerance of ±0.05. The DF of 0.0037 at 10 GHz indicates the material's low loss at higher frequencies, while a TCDk of +50 ppm/°C from −50 to +150°C shows how it maintains constant Dk over a wide temperature range. It has a low moisture absorption value of 0.06%. de

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ROGERS CORP., Advanced Connectivity Solutions, 100 S. Roosevelt Ave., Chandler, AZ 85226-3415; (480) 961-1382, 1-877-643-7701, FAX: (480) 961-4533, www.rogerscorp.com

(Continued from page 46)

Unmanned aerial vehicles (UAVs) have been used for some time as information-gathering surveillance tools and even as the means of delivering explosive devices to remote positions.

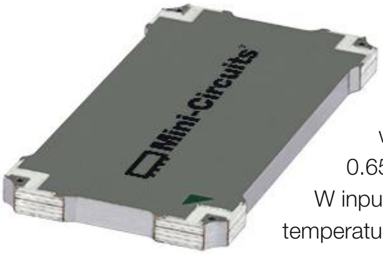
More advanced, humanoid-type robots could take up a call to duty alongside human partners, with many advantages, including being immune to chemical and biological weapons and being free of the "decision-making clutter" brought about by having emotions. Unlike humans, they can be readily repaired when damaged on the battlefield and be equipped with multiple sensors to far exceed the sensory capabilities of a human soldier. But might the inclusion of emotions bring benefits to a robotic warrior?

Human soldiers learn to count on each other in tight spots, often surpassing normal physical capabilities with a rush of adrenaline to save a comrade under stress or rescue the member of a troop who has become more like family than simply a group of warriors with a common goal. Key decisions on the battlefield are still made by humans even as they rely more on robotic systems such as UAVs as part of an assignment.

Robots can be programmed with impressive precision and positional capabilities through a combination of digital hardware and software, as well as the ability to make fundamental decisions according to a set of predictable situations. But when the number of variables exceeds the limit of a robot's programming, either a compromise decision or no decision will be made in response to a situation. In that case, the infinite decision-making capabilities of a human commander are required.

Robots offer the capabilities to make a battlefield safer for humans, and they can add logic to a normally emotionally charged environment. Adding programmed emotions might even increase the value of battlefield robotic systems by providing a greater range of decision-making when needed. After all, in a crisis, even Mr. Spock cast aside logic in favor of the emotions bestowed upon him by a human mother. de

Robots offer the capabilities to make a battlefield safer for humans, and they can add logic to a normally emotionally charged environment.

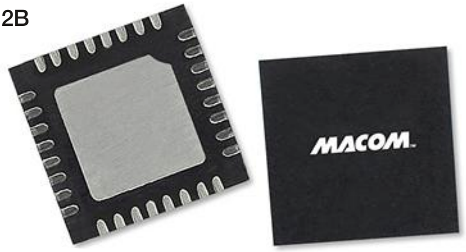


Surface-Mount Coupler Handles 100 W to 3 GHz

MINI-CIRCUITS MODEL BDCH-15-33+ is a bidirectional surface-mount bidirectional coupler designed for high input power from 0.5 to 3.0 GHz. With worst-case insertion loss of 0.45 dB (typically 0.25 dB) and thermal resistance of typically 0.65°C/W across the full frequency range, the compact coupler handles as much as 100 W input power at +85°C case temperature and as much as 70 W input power at +105°C case temperature. Return loss is typically 30 dB at all ports. The RoHS-compliant coupler provides dc-pass capability from input to output ports. With typical directivity of 25 dB and typical coupling of 15.5 ±1 dB, the bidirectional coupler is a good fit for connecting power amplifiers with distributed antenna networks. It comes in a compact housing measuring 1.000 × 0.500 × 0.051 in. (25.40 × 12.70 × 1.30 mm) with wraparound terminations and is rated for operating case temperatures of –55 to +105°C.

MINI-CIRCUITS

P.O. Box 350166, Brooklyn, NY 11235-0003; (718) 934-4500,
https://www.minicircuits.com/WebStore/dashboard.html?model=BDCH-15-33%2B



Distributed Amplifier Ranges 2 to 18 GHz

MODEL MAAL-011151 is a low-noise distributed amplifier in a 32-lead PQFN package for applications from 2 to 18 GHz. It provides 15-dB linear gain across its bandwidth with noise figure of 3.5 dB at 10 GHz. The output power at 1-dB compression is typically +19 dBm, while the return loss is 10 dB or better across the frequency range. The RF input and output ports, which are impedance matched to 50 Ω, are dc blocked. The RoHS-compliant GaAs heterojunction-bipolar-transistor (HBT) amplifier is well-suited for use in test-and-measurement, EW, and ECM systems.

MACOM TECHNOLOGY SOLUTIONS INC.

100 Chelmsford St., Lowell, MA 01851; (978) 656-2896, (800) 366-2266, www.macom.com

Directional Coupler Covers 0.5 to 26.5 GHz

MODEL C-0526-10 is a 10-dB stripline directional coupler for use from 0.5 to 26.5 GHz. It maintains 10-dB coupling across the wide frequency range with coupling flatness within ±1 dB. The maximum insertion loss is 2.2 dB while the maximum input and output VSWR is 1.40:1. The directional coupler, which handles average power levels as high as 20 W and peak power levels to 3 kW, achieves minimum directivity of 14 dB. It measures 2.90 × 0.650 × 0.250 in. and is equipped with female SMA connectors. The C-0526-10 is designed for operating temperatures from –32 to +85°C.

ELECTROMAGNETIC TECHNOLOGIES INDUSTRIES INC. (ET INDUSTRIES)

50 Intervale Road, Boonton, NJ, 07005; sales@etiworld.com, PH: 001+(973)-394-1719, FAX: 001+(973) 394-1710, www.etiworld.com

Limiter Line Grows with Coverage to 40 GHz

A LINE OF HIGH-POWER coaxial limiters has been expanded to include 13 models covering a total frequency range of 200 MHz to 40 GHz. Well-suited for communications, radar, and electronic-warfare (EW) applications, the limiters provide high suppression of even-order harmonics with good leakage protection at power levels to +18 dBm. The rugged 50-Ω limiters can handle peak power levels to 200 W with fast recovery times of 10 to 100 ns and can be used without external impedance-matching components. They are designed to meet MIL-STD-202 environmental conditions for shock, humidity, vibration, altitude, and temperature cycling. The limiters, supplied in compact housings with field-replaceable connectors, are designed for an operating temperature range of –54 to +85°C.

PASTERNACK ENTERPRISES INC.

P. O. Box 16759, Irvine, CA 92614; (949) 261-1920, FAX: (949) 261-7451, sales@pasternack.com, www.pasternack.com

CHRIS DeMARTINO | Editor

Foundry Aims to Serve All Corners

In this Q&A, Amol Kalburge, senior director of RF and high-performance analog marketing and strategic accounts at TowerJazz, discusses the technologies and different markets that define his company.

Can you first give us an overview of TowerJazz?

TowerJazz *(www.towerjazz.com)* is the global specialty analog foundry leader. We manufacture next-generation integrated circuits (ICs) in growing markets such as 5G/IoT, optical networking, cloud computing, automotive ADAS and vehicle electrification, high-performance imaging including AR/VR, aerospace and defense, and smart power management. TowerJazz offers a broad range of customizable process platforms such as silicon-germanium (SiGe) BiCMOS, RF silicon-on-insulator (SOI), RF CMOS, mixed-signal/CMOS, sensors including CMOS image sensors and non-imaging sensors, integrated power management, and MEMS.

TowerJazz also provides world-class design enablement for a quick and accurate design cycle as well as transfer optimization and development process services (TOPS) to integrated device manufacturers (IDMs) and fabless companies that need to expand capacity. To provide multi-fab sourcing and extended capacity for its customers, TowerJazz operates two manufacturing facilities in Israel (150 mm and 200 mm), two in the U.S. (200 mm), and three facilities in Japan (two 200 mm and one 300 mm). All our fabs are certified for IATF 16949, the industry's highest standard of quality system for automotive manufacturing *(Fig. 1)*.

What are some of the technologies the company is involved with?

We offer a broad range of advanced specialty analog technologies, with technology nodes ranging from > 0.25 µm

Migdal Haemek, Israel
- **6-inch** (150mm)
- CMOS, CIS, Power, Power Discrete
- 1µm to 0.35µm
- Planarized BEOL, W and Oxide CMP

Migdal Haemek, Israel
- **8-inch** (200mm)
- CMOS, CIS, Power, Power Discrete, RF Analog, MEMS
- 0.18µm to 0.13µm
- Cu and Al BEOL, EPI, 193nm Scanner

Newport Beach, CA, USA
- 8 inch (200mm)
- CMOS, CIS, RF Analog, MEMS
- 0.18µm to 0.13µm
- Al BEOL, SiGe, EPI

San Antonio, TX, USA
- 8-inch (200mm)
- Power, RF Analog
- 0.18µm
- Al BEOL

Arai, Japan
- 8-inch (200mm)
- Analog, CIS
- 0.13µm to 0.11µm

Tonami, Japan
- 8-inch (200mm)
- Analog, Power Discrete, NVM, CCD
- 0.35µm to 0.15µm

Uozu, Japan
- 12-inch (300mm)
- Analog, CMOS, CIS, RFCMOS/SOI
- 65nm & 45nm

1. TowerJazz has facilities located in multiple spots around the world.

down to 45 nm. I would urge readers to visit our website for details on all of the technologies we offer, but I will highlight some of the firsts we have brought to market in the arena of microwaves and RF. Two decades ago, we were the first pure-play foundry to offer advanced SiGe technologies for wireless and wireline markets. Today, we are in production with our fifth-generation SiGe technology with an Ft/Fmax of greater than 300 GHz. In addition, we are working on sixth-generation to usher in the era of terabit optical communications.

We were also the first pure-play foundry to offer advanced RF SOI technology back almost a decade ago to enable advanced front-end modules for 3G/4G-enabled smartphones and are now in production with our fourth-generation RF SOI technology. We are continuing to push the Ron-Coff—a key figure of merit for an RF switch—to below 50 fs and plan to release newer generations in coming years.

And finally, last year we started offering the industry's first open silicon photonics (SiPho) process and process design kits (PDKs) for growing optical networking and data center markets. This open SiPho foundry technology is also rapidly becoming a technology of choice for the new and exciting market of LiDAR to make autonomous driving a reality. There are many more firsts across TowerJazz that we are proud of, and we continue to break new ground as we pursue most aggressive technology roadmap to help our customer partners bring unique and exciting new products to market rapidly.

Can you talk about some of the most difficult challenges you've faced in recent years?

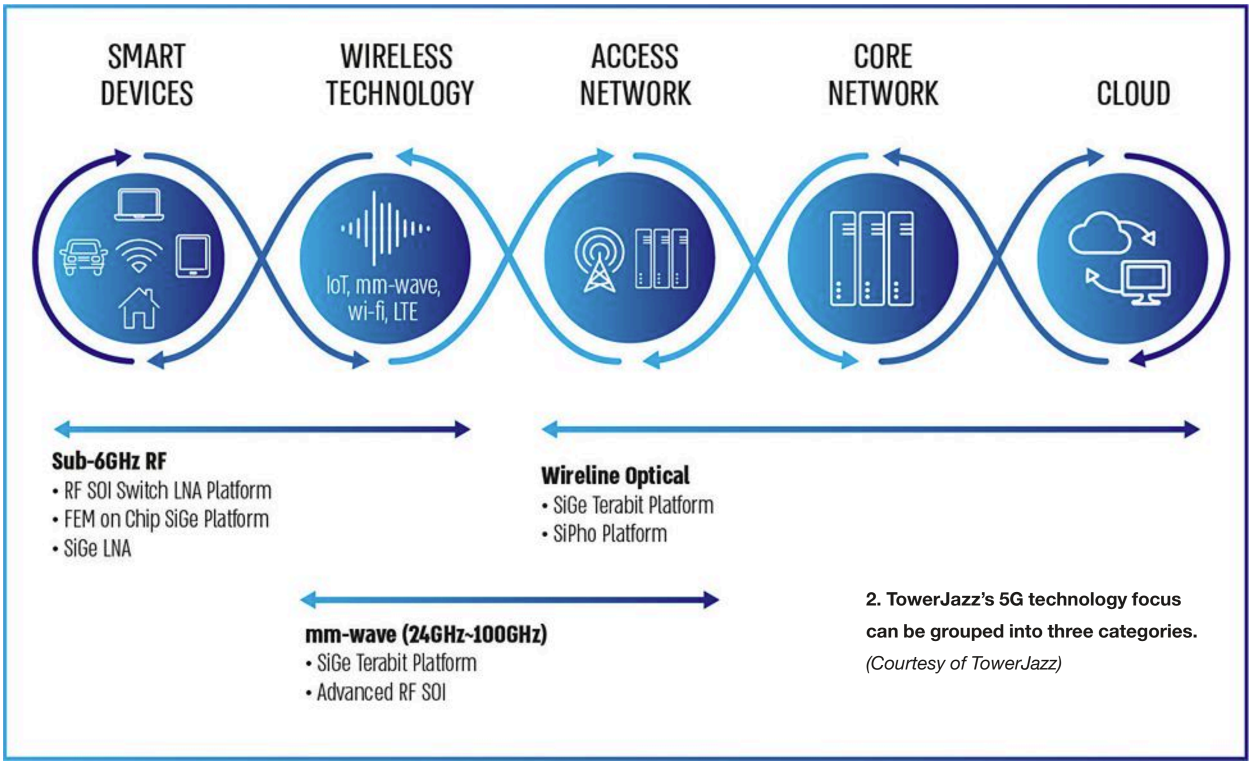
It's an interesting question and perhaps not easily answerable. Challenges always feel insurmountable when they are in front of you. But once you overcome them, they don't feel as difficult in the rear mirror. As I described earlier, TowerJazz prides itself in being, or becoming, the first foundry to provide the most advanced solutions in everything we do. Being first means being unafraid of taking on new challenges and learning from history. We have been able to overcome challenges by being committed to our values, being loyal to our customers, and most importantly, continuing to invest in people to pursue excellence—be it a new technology, lean and efficient manufacturing, or quality.

Tell us about some of the markets that the end products are serving.

Our specialty analog solutions are aligned with three global megatrends:

1. Wireless everywhere (seamless connectivity)
2. Green everything (or energy efficiency)
3. Smart everything (or intelligent systems)

Our CMOS image sensor technology is recognized by many market leaders as best in class. We enjoy a leading market share in high-end photography and cinematography, industrial/machine vision, x-ray imaging, ADAS and augmented/virtual reality (AR/VR), and with our latest 65-nm smallest-in-the-world global



2. TowerJazz's 5G technology focus can be grouped into three categories.
(Courtesy of TowerJazz)

shutter technology, we are raising the bar even higher. Our advanced power-management platform is a technology of choice for leading solutions providers for electric-vehicle battery-management systems, data-center power supplies, industrial motor control, and LED lighting, to name a few.

In the RF/microwave arena, we are serving the entire value chain starting with cellular/Wi-Fi/IoT on the fronthaul to mmWave/microwave backhaul, to optical fiber networks. Some notable examples of end products are smartphone front-end modules (power amplifiers, RF antenna switches and tuners, and low-noise amplifiers), optical network elements (transimpedance amplifiers), clock-data recovery, laser drivers, compute and storage ICs (line drivers), HDD preamps, and analog-to-digital converters (ADCs).

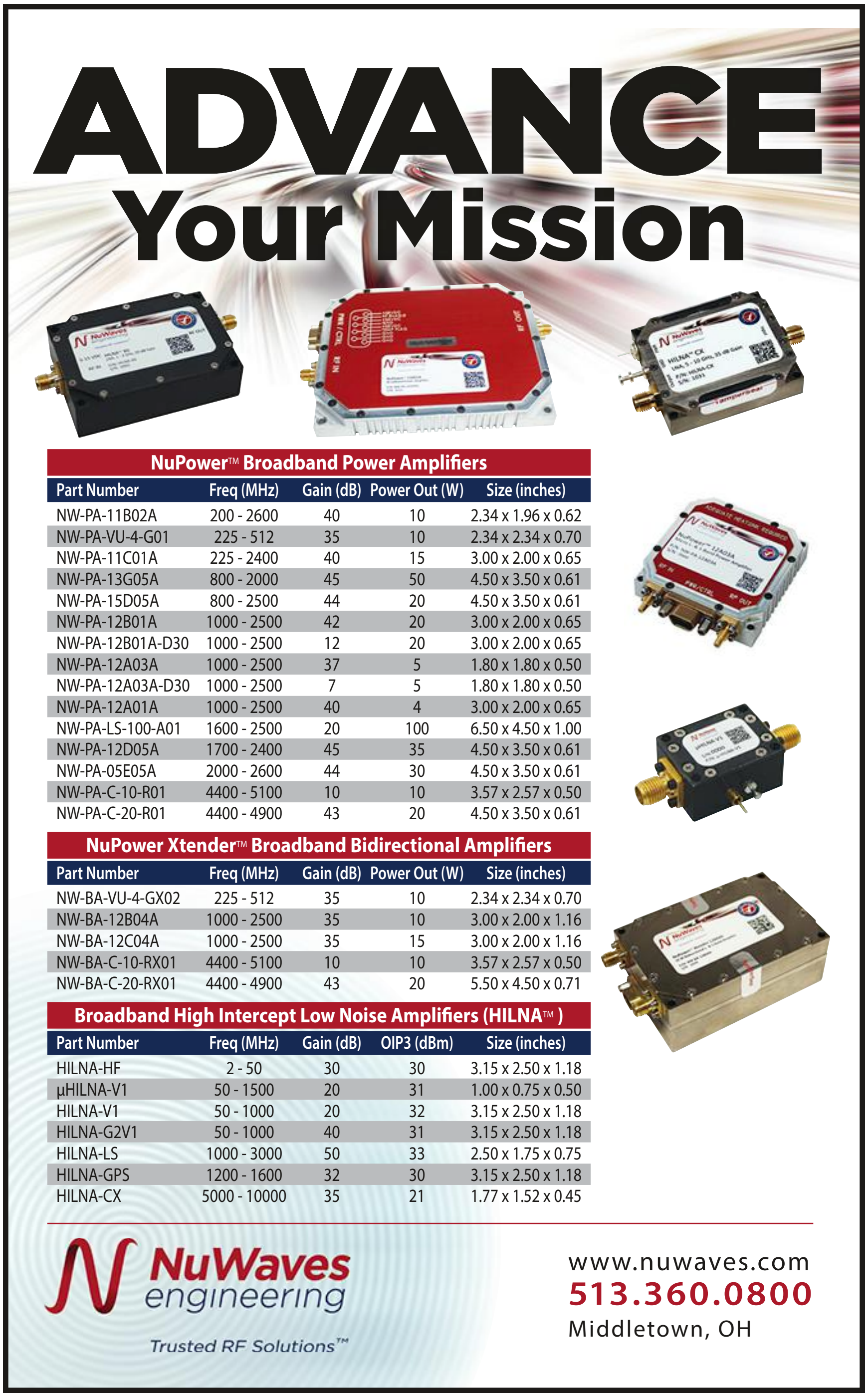
Finally, we also serve the growing market of automotive radar and silicon photonics. Our silicon photonics platform is being adopted rapidly to meet worldwide data-traffic demands that 5G, cloud computing, and IoT are driving.

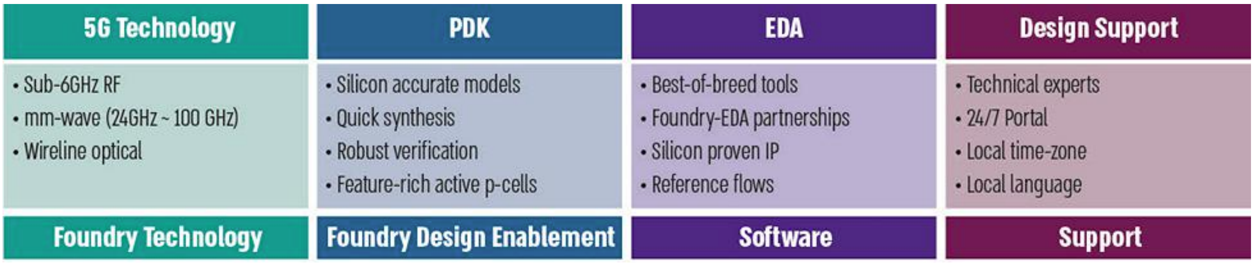
Also, I would be remiss if I didn't mention a host of smart-sensor applications that we serve with our non-imaging sensor technologies, such as hazardous gas sensors, temperature and humidity sensors, and radiation sensors.

With 5G being all the rage, what role do you see TowerJazz playing in terms of 5G?

We do believe that 5G technology will be a paradigm shift in how we and the "things" communicate with each other. 5G is about more than faster upload/download speeds. 5G networks will support massive and universal machine-to-machine communication, connecting billions of sensors and machines securely, reliably, and with ultra-low latency. On the backhaul, telecom operators will not only require higher-speed fixed wireless and fiber-optic networks, but also sophisticated integration of ubiquitous computing and storage architectures.

For multiple generations of network architectures, our SiGe, SOI, RFCMOS, and MEMS technologies have been extensively used by market leaders to design building blocks for the entire value chain. This multi-generational partnership and roadmap alignment with market leaders gives us unique insights into what is needed from silicon technology—or any other semiconductor tech for that matter—to support grueling performance demands of future 5G networks. We have significantly enhanced—and customized—our proven technology platforms and are





5G Technology	PDK	EDA	Design Support
• Sub-6GHz RF • mm-wave (24GHz ~ 100 GHz) • Wireline optical	• Silicon accurate models • Quick synthesis • Robust verification • Feature-rich active p-cells	• Best-of-breed tools • Foundry-EDA partnerships • Silicon proven IP • Reference flows	• Technical experts • 24/7 Portal • Local time-zone • Local language
Foundry Technology	Foundry Design Enablement	Software	Support

3. From 5G to EDA, TowerJazz brings a range of capabilities to the table.

collaborating with our partners to help them bring exciting next-generation 5G products to market.

We group our 5G technology platforms into three broad categories *(Fig. 2)* depending on the class of network connectivity:

1. Sub-6-GHz RF
2. Millimeter-wave (mmWave)
3. Wireline optical

I would urge readers to learn more details about these platforms on our website or at one of our technology symposiums.

We pride ourselves in bringing not only leading-edge silicon technologies to the market, but also being best-in-class in terms of design enablement and design support. We believe that no matter how good a technology, it's of no use if our customers cannot bring products to market in the shortest possible time. Our PDKs, silicon-proven models, and IP allow our customers to design their products once and design them right. Our holistic approach to technology delivery is what differentiates us from other foundries. This is why we have been the largest specialty analog foundry for almost a decade now *(Fig. 3)*.

Here are some useful links for readers to learn more about our technologies:

https://towerjazz.com/technology/process-technology-offerings

https://towerjazz.com/2019/05/30/gearing-up-for-5g-with-towerjazz

https://towerjazz.com/news-events/events **mw**

CHRIS DeMARTINO | Editor

Workflow Carves a Path to EFFECTIVE FILTER DESIGN

A simulation-based filter design example demonstrates a semi-automated workflow that can help designers achieve first-pass success.

Simulation-based filter design can be tackled in different ways thanks to the various simulation tools available and the functionality within them. Generally, these tools offer a range of features that help build a more streamlined design process for both seasoned designers and novices alike. This article presents a workflow for designing microstrip filters that leverages two software tools in tandem: FilterSolutions from Nuhertz Technologies' *(www.nuhertz.com)* along with Sonnet Software's *(www.sonnetsoftware.com)* electromagnetic (EM) simulation engine, known as em.

With Nuhertz's FilterSolutions software, users can design various types of filters, such as lumped-element and distributed versions. One highly effective aspect of FilterSolutions is that it works in conjunction with various third-party EM simulators. As a result, a filter that's created in FilterSolutions can then be exported to an EM simulator of one's choice for simulation and analysis.

Among the EM simulators that work well with FilterSolutions is Sonnet's EM analysis tool. With a history of over 35 years, Sonnet prides itself on its long track record of "accuracy and innovation" in the world of EM simulation. Sonnet can be used for a variety of use cases, such as integrated-circuit (IC) design (RFICs and MMICs), low-temperature co-fired ceramics (LTCC) and thin-film design, as well as printed-circuit-board design (single- and multi-layer laminates).

In the example presented here, a microstrip interdigital bandpass filter will be created in FilterSolutions and then exported to Sonnet. The exported filter incorporates parameterized Sonnet geometry, which can greatly help in terms of optimizing filter designs to achieve the desired performance.

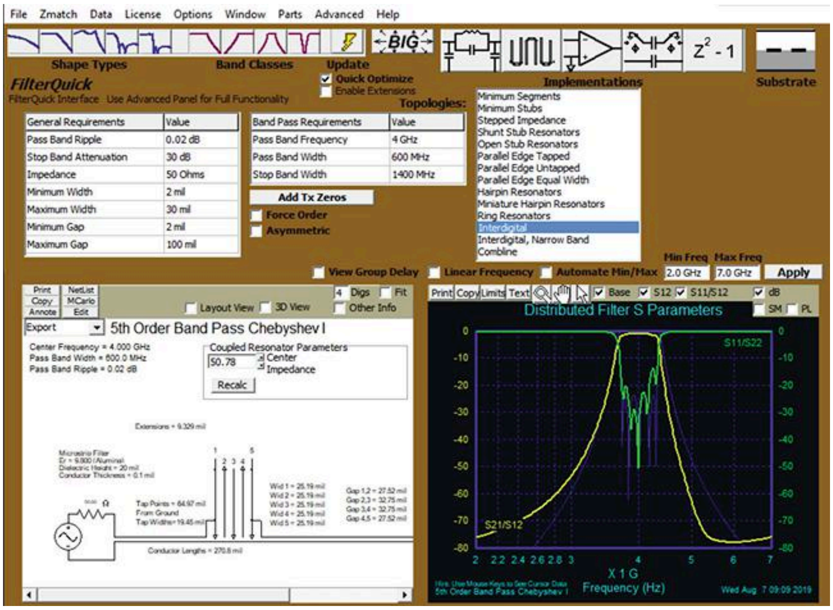
STARTING THE PROCESS IN FILTERSOLUTIONS

The design process begins by first entering the desired parameters into FilterSolutions *(Fig. 1)*. It's worth noting that FilterSolutions offers the option to select either an advanced user design panel or a simplified version, known as *FilterQuick*. In this example, *FilterQuick* is used.

As seen in *Figure 1*, a microstrip interdigital bandpass filter with a center frequency of 4 GHz will be designed. The filter is a fifth-order Chebyshev implementation. A 20-mil-thick alumina substrate was chosen.

Simulated S-parameters of the filter are shown on the lower right of *Figure 1*. The minimum and maximum frequencies are set to 2 GHz and 7 GHz, respectively (seen directly above the S-parameter plots). This frequency range will automatically be set as the swept analysis range in Sonnet after the filter is exported.

Before exporting the filter to Sonnet, some steps can be taken to allow for a more efficient analysis. Clicking on the simplified filter schematic on the left side of *Figure 1* reveals several modifiable parameters *(Fig. 2)*. The parameters are denoted as *Length*



☑ Quick Optimize
☐ Enable Extensions

General Requirements	Value
Pass Band Ripple	0.02 dB
Stop Band Attenuation	30 dB
Impedance	50 Ohms
Minimum Width	2 mil
Maximum Width	30 mil
Minimum Gap	2 mil
Maximum Gap	100 mil

Band Pass Requirements	Value
Pass Band Frequency	4 GHz
Pass Band Width	600 MHz
Stop Band Width	1400 MHz

☐ Force Order
☐ Asymmetric

Minimum Segments
Minimum Stubs
Stepped Impedance
Shunt Stub Resonators
Open Stub Resonators
Parallel Edge Tapped
Parallel Edge Untapped
Parallel Edge Equal Width
Hairpin Resonators
Miniature Hairpin Resonators
Ring Resonators
Interdigital
Interdigital, Narrow Band
Combline

☐ View Group Delay ☐ Linear Frequency ☐ Automate Min/Max
Min Freq: 2.0 GHz
Max Freq: 7.0 GHz

4 Digs ☐ Fit
☐ Layout View ☐ 3D View ☐ Other Info
☑ Base ☑ S12 ☑ S11/S12 ☑ dB
☐ SM ☐ PL

5th Order Band Pass Chebyshev I

Center Frequency = 4.000 GHz
Pass Band Width = 600.0 MHz
Pass Band Ripple = 0.02 dB

Coupled Resonator Parameters
50.78 Center Impedance

Extensions = 9.329 mil

Microstrip Filter
Er = 9.800 (Alumina)
Dielectric Height = 20 mil
Conductor Thickness = 0.1 mil

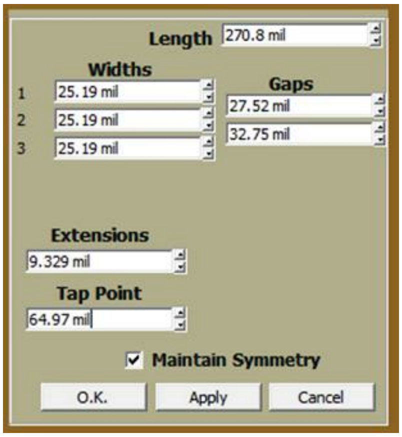
Tap Points = 64.97 mil
From Ground
Tap Widths= 19.45 mil

Wid 1 = 25.19 mil
Wid 2 = 25.19 mil
Wid 3 = 25.19 mil
Wid 4 = 25.19 mil
Wid 5 = 25.19 mil

Gap 1,2 = 27.52 mil
Gap 2,3 = 32.75 mil
Gap 3,4 = 32.75 mil
Gap 4,5 = 27.52 mil

Conductor Lengths = 270.8 mil

1. Shown are the parameters for this fifth-order filter implementation along with the simulated S-parameters.



Length 270.8 mil

Widths
1 25.19 mil
2 25.19 mil
3 25.19 mil

Gaps
27.52 mil
32.75 mil

Extensions
9.329 mil

Tap Point
64.97 mil

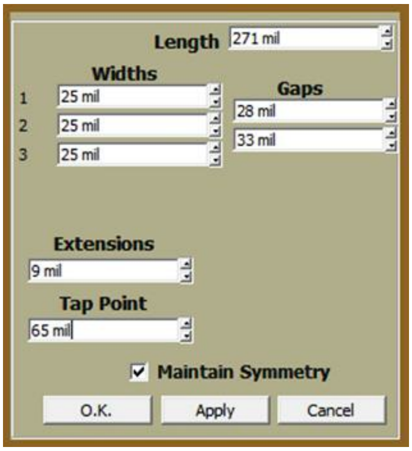
☑ Maintain Symmetry

2. These parameters can be adjusted by the user.

(length of the resonators), *Widths* (widths of the resonators), *Gaps* (gaps between resonators), *Extensions* (length of the resonator extensions), and *Tap Point* (position of the input/output feed lines).

The parameters described will transfer over to Sonnet following exportation. One aspect to keep in mind, though, is that Sonnet utilizes a grid. When performing a simulation, Sonnet will analyze the metallization that's fit to the user-selectable grid (known as the cell size). In this case, the filter will be analyzed in Sonnet with the cell size set to 1 × 1 mil. Hence, an effective approach is to round the parameters just mentioned to the nearest mil before exporting the filter to Sonnet so that the filter dimensions will correlate with the Sonnet grid. *Figure 3* shows the new parameter values after being rounded.

Now that the process of creating the filter in FilterSolutions is complete, the next step is to export the filter to Sonnet for EM analysis. Those with experience designing RF/microwave filters surely know that EM analysis is critical to accurately predict the performance of a filter like the one being designed here. That's because, when modeling distributed filters, circuit simulators



Length 271 mil

Widths
1 25 mil
2 25 mil
3 25 mil

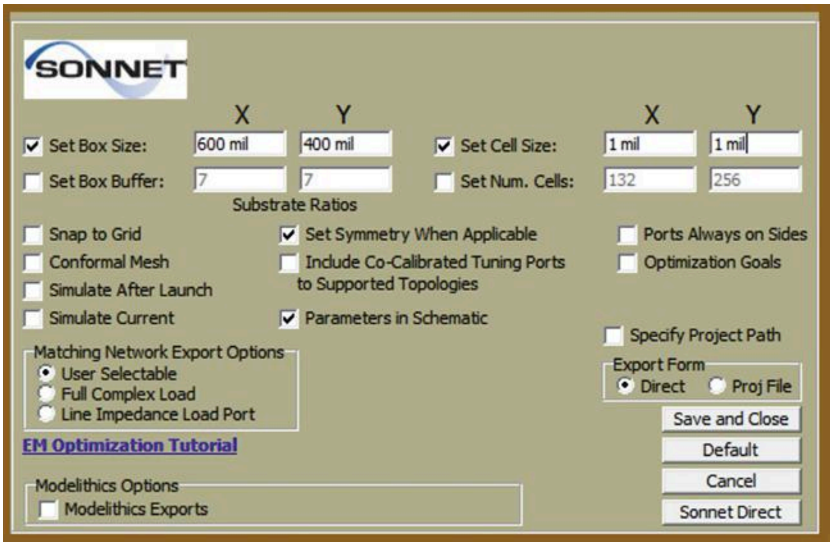
Gaps
28 mil
33 mil

Extensions
9 mil

Tap Point
65 mil

☑ Maintain Symmetry

3. Shown are the values of the parameters after being rounded to the nearest mil.



	X	Y
☑ Set Box Size:	600 mil	400 mil
☐ Set Box Buffer:	7	7

	X	Y
☑ Set Cell Size:	1 mil	1 mil
☐ Set Num. Cells:	132	256

☐ Snap to Grid
☐ Conformal Mesh
☐ Simulate After Launch
☐ Simulate Current

☑ Set Symmetry When Applicable
☐ Include Co-Calibrated Tuning Ports to Supported Topologies
☑ Parameters in Schematic

☐ Ports Always on Sides
☐ Optimization Goals

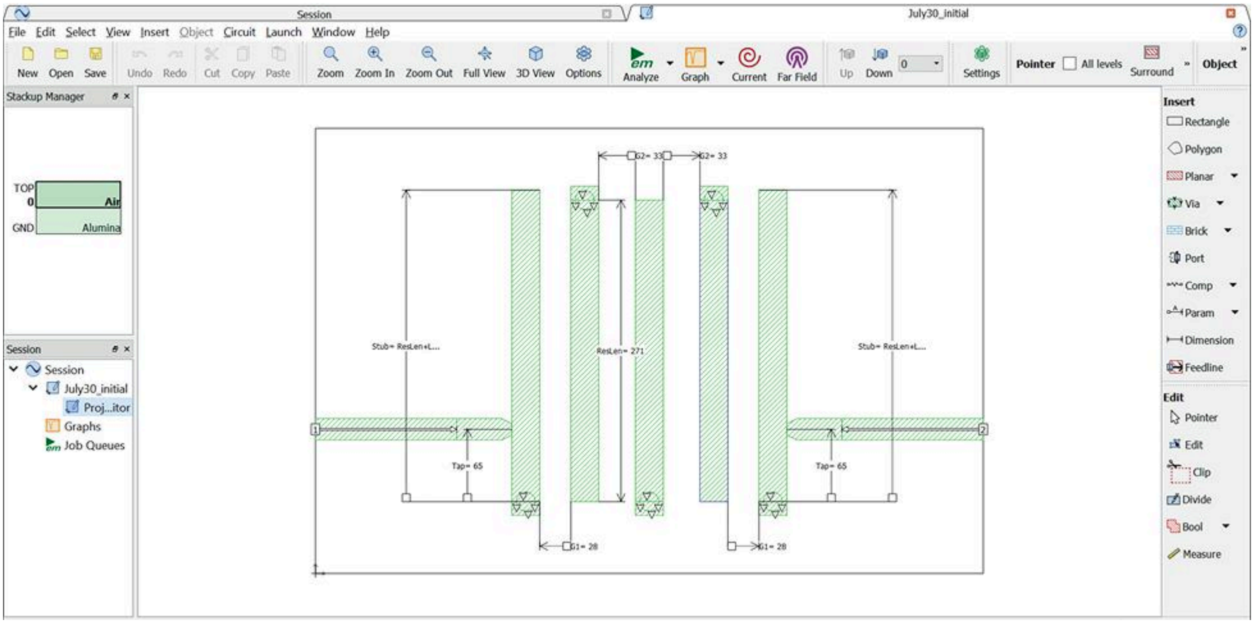
☐ Specify Project Path

Matching Network Export Options
● User Selectable
○ Full Complex Load
○ Line Impedance Load Port

Export Form
● Direct ○ Proj File

Modelithics Options
☐ Modelithics Exports

4. Users have various options to choose from when exporting a design to Sonnet.



5. The Sonnet project includes parameterized elements like *ResLen* and *Tap*.

have limited accuracy due to factors such as cross-element coupling, housing effects, evanescent modes, etc. Designers must therefore utilize an EM simulator to analyze and then optimize the design. However, performing a direct EM optimization to adjust a filter's physical dimensions can be rather time-consuming.

Nuhertz emphasizes that FilterSolutions allows for an accurate starting point for EM optimization, thereby minimizing the time required. In other words, after a filter is exported from FilterSolutions to an EM simulator, the results of the EM analysis should be close to the simulated results generated by FilterSolutions.

What this ultimately means is that the initial EM analysis results should come reasonably close to the desired goals. As a result, filters can be optimized, or tuned, with a minimal number of EM simulation runs, making it possible to achieve the final performance goals more quickly. Hence, the design process is more efficient. This capability will be revealed later.

As a final step before exporting, a touchstone file was saved that contains the S-parameter data generated by FilterSolutions. This data will later be compared to the simulated results in Sonnet. To create a touchstone file, one must simply select *Touchstone S* from the *Export* dropdown menu and then save the file.

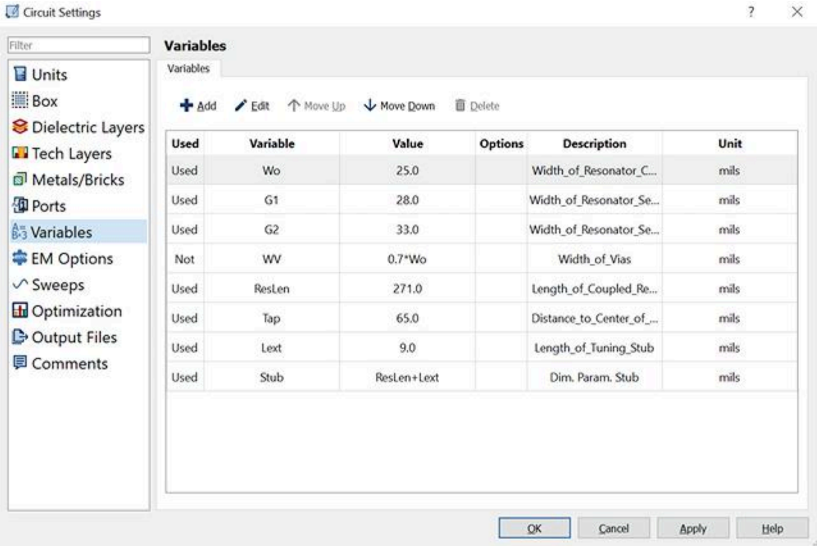
FilterSolutions allows for an accurate starting point for EM optimization, thereby minimizing the time required. In other words, after a filter is exported from FilterSolutions to an EM simulator, the results of the EM analysis should be close to the simulated results generated by FilterSolutions.

EXPORTING TO SONNET

Now, the filter can be exported to Sonnet by selecting *Sonnet Setup* from the *Export* dropdown menu. *Figure 4* reveals the subsequent user interface. Note that the cell size is set to 1 × 1 mil. In addition, by selecting the *Parameters in Schematic* checkbox, the Sonnet project will incorporate parameterized geometry.

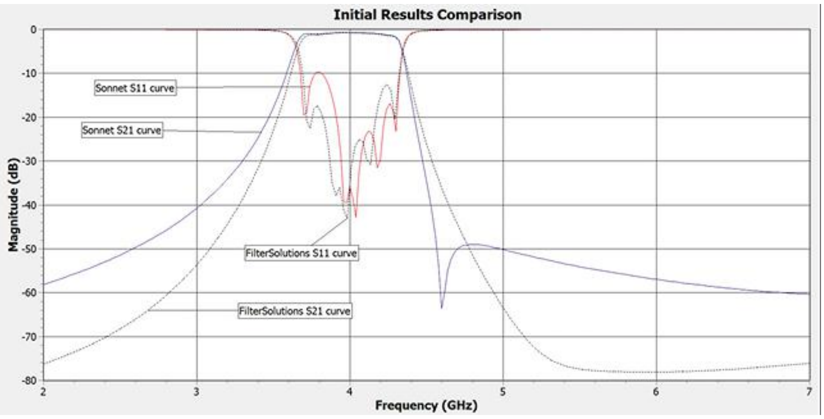
Clicking *Sonnet Direct (Fig. 4, again)* automatically creates the filter. *Figure 5* shows the design in Sonnet's new and enhanced version 17 interface. Note that various filter elements, such as the length of the resonators, the spacing between resonators, and the tap points, are parameterized due to enabling the *Parameters in Schematic* option in FilterSolutions.

By clicking *Settings* in the ribbon bar and then selecting *Variables*, users can view the parameterized elements in tabular form *(Fig. 6)*. Shown are *G1* and *G2*, which denote the amount of separation, or gaps, between resonators. These variables correspond to the *Gaps* parameters in FilterSolutions *(Fig. 3, again)*. The *ResLen* variable, which corresponds to the FilterSolutions *Length* parameter, represents the length of the resonators.

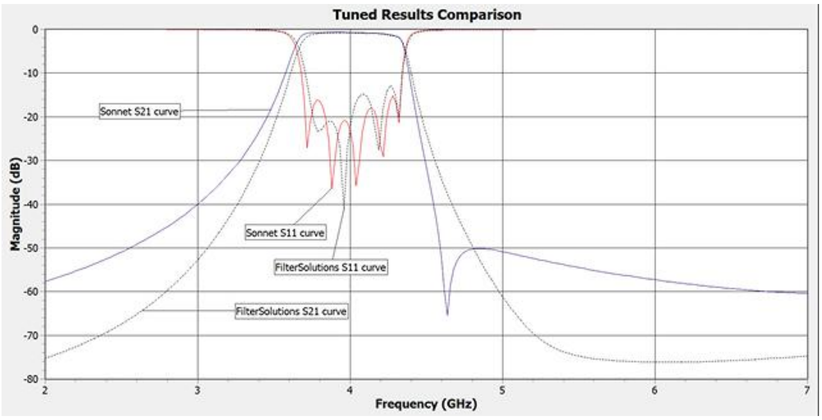


Used	Variable	Value	Options	Description	Unit
Used	Wo	25.0		Width_of_Resonator_C...	mils
Used	G1	28.0		Width_of_Resonator_Se...	mils
Used	G2	33.0		Width_of_Resonator_Se...	mils
Not	WV	0.7*Wo		Width_of_Vias	mils
Used	ResLen	271.0		Length_of_Coupled_Re...	mils
Used	Tap	65.0		Distance_to_Center_of_...	mils
Used	Lext	9.0		Length_of_Tuning_Stub	mils
Used	Stub	ResLen+Lext		Dim. Param. Stub	mils

6. All of the variables are listed here in a tabular format.



7. These are the simulated S-parameter results of the initial Sonnet analysis, along with the FilterSolutions results.



8. After applying adjustments made in FilterSolutions, the Sonnet analysis provided these simulated S-parameter results. The filter now achieves desirable performance.

The *Tap* and *Lext* variables equate to the *Tap Point* and *Extensions* parameters in FilterSolutions, respectively. Finally, *Stub* is simply *ResLen* plus *Lext*.

Figure 7 shows the results of the Sonnet simulation, with the blue and red traces representing the simulated S_{21} and S_{11}, respectively. For comparison purposes, *Figure 7* also reveals the FilterSolutions simulation results, with both S_{21} and S_{11} represented by dashed black traces.

FilterSolutions did provide a very good starting point—the FilterSolutions simulation results are reasonably close to the results of the Sonnet simulation. Nevertheless, as one might expect, the filter needs to be optimized, since the performance does not quite meet the goals. Specifically, S_{11} in the low end of the passband needs to improve.

OPTIMIZING THE FILTER

Design optimization can be accomplished in several different ways. One method is to perform a direct EM optimization, which is included in Sonnet. Another more manual approach involves returning to FilterSolutions and adjusting the modifiable parameters that were shown earlier. One can simply adjust any or all of these parameters and then immediately see how the S-parameters change in FilterSolutions.

In this example, the primary objective would be to adjust the parameters to improve S_{11} toward the low end of the passband (the goal is for S_{11} to stay below −15 dB throughout the entire passband). Once the new value(s) are determined, the change(s) can easily be applied to the Sonnet project thanks to the parameterized geometry. In this case, such an approach will likely require several Sonnet analyses, but it should take a relatively short amount of time.

Another approach for optimization is the port-tuning method, which involves placing internal ports at strategic locations in the filter layout. Tuning elements are inserted at the port locations to determine the required layout adjustments. The port-tuning method can be performed with any circuit simulator that offers capacitor, inductor, and transmission-line elements, such as Sonnet's netlist-engine feature.

For this example, the filter will be manually tuned using FilterSolutions. The port-tuning method will be the subject of a future article.

The tunable FilterSolutions parameters were shown in *Figure 3*. It was determined that adjusting the *Extensions* parameters noticeably affected the filter's S_{11} performance. After a bit of tuning, the final value for the *Extensions* parameter (9 mils was the original value) was ascertained to be 12 mils. In addition, slight changes were made to the *Length* parameter (changed from 271 to 270 mils) and the first *Gaps* parameter (changed from 28 to 27 mils).

Before returning to Sonnet, a new touchstone file was saved that contains the updated S-parameters. Upon return to Sonnet, the variables can be updated by clicking *Settings* in the ribbon bar and then selecting *Variables (Fig. 6, again)*.

Figure 8 reveals the results after performing a new simulation in Sonnet. Again, the blue and red traces represent the simulated S_{21} and S_{11}, respectively.

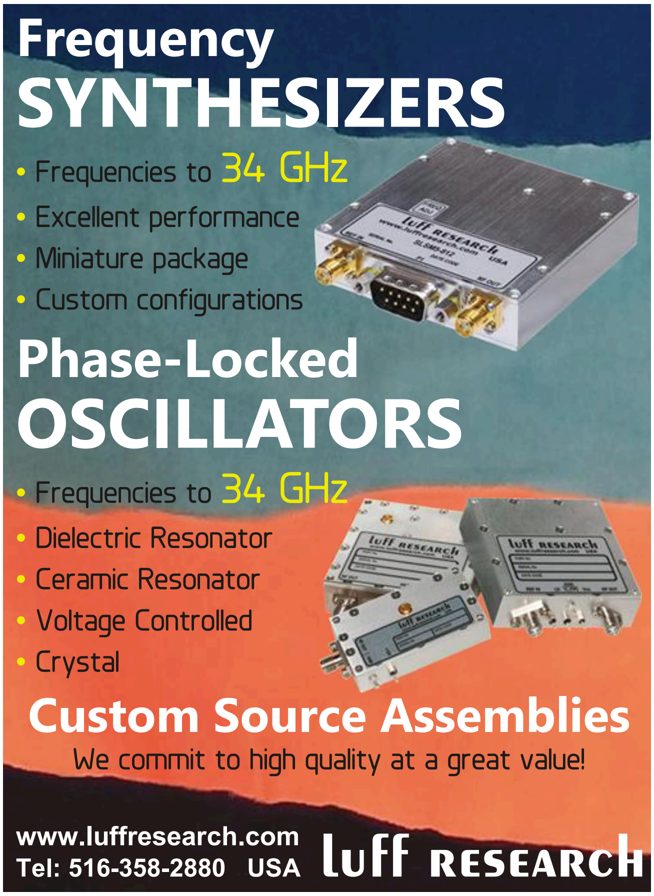
The Nuhertz-Sonnet combo is a proven, productive way to achieve successful filter designs.

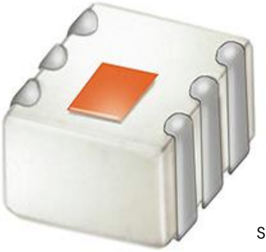
In addition, the FilterSolutions simulation results (S_{21} and S_{11}) of the tuned filter are once again represented by dashed black traces for comparison purposes.

The results of the Sonnet simulation reveal good performance. Thus, the design process is complete. The total time required to complete the entire process was approximately a few hours, with most of that time spent tuning the design. The actual EM analysis time for this filter was only one to two minutes on a desktop machine.

In closing, leveraging the combination of Nuhertz's FilterSolutions and Sonnet can be an effective way to design RF/microwave filters. FilterSolutions offers an easy-to-use interface that even novices should be able to grasp. Sonnet is another user-friendly tool with a reputation as an extremely accurate EM simulator. With features like parameterized Sonnet geometry, the Nuhertz-Sonnet combo is a proven, productive way to achieve successful filter designs. mw

ACKNOWLEDGEMENTS: I would like to thank Jeff Kahler from Nuhertz and Greg Kinnetz and Brian Rautio from Sonnet for supporting this article.





10-dB LTCC Coupler Channels 2.4 to 2.5 GHz

MINI-CIRCUITS' MODEL CPJC-10-252R+ is a low-temperature-co-fired-ceramic (LTCC) directional coupler for applications from 2.4 to 2.5 GHz. It delivers typical directivity of 19 dB (and minimum of 11 dB) while maintaining 10-dB coupling flat within ±1 dB across the frequency range. The mainline insertion loss is typically 0.65 dB and no more than 1.00 dB, while typical input and output return loss is 17 dB. The RoHS-compliant 10-dB coupler is ideal for applications at industrial-scientific-medical (ISM) band frequencies, including in Bluetooth, WLAN, and Zigbee systems. It's supplied in a miniature 0603 type case measuring 0.063 × 0.031 × 0.024 in. (1.60 × 0.79 × 0.61 mm). The 50-Ω LTCC coupler is designed for operating temperatures from –40 to +85°C.

MINI-CIRCUITS, P.O. Box 350166, Brooklyn, NY 11235-0003; (718) 934-4500,
https://www.minicircuits.com/WebStore/dashboard.html?model=CPJC-10-252R%2B

Low-Noise Amplifier Gains 5 to 11 GHz

MINI-CIRCUITS' MODEL ZX60-05113LN+ is a rugged low-noise amplifier (LNA) with high gain and low noise figure from 5 to 11 GHz. It provides 22-dB typical gain with ±0.7-dB gain flatness across the full bandwidth with typical noise figure of 2.3 dB from 5 to 7 GHz, 1.8 dB from 7 to 9 GHz, and 1.7 dB from 9 to 11 GHz. The input return loss is typically 6.7 dB from 5 to 7 GHz, 12.1 dB from 7 to 9 GHz, and 9.0 dB from 9 to 11 GHz. The output return loss is typically 12.4 dB from 5 to 7 GHz, 17.0 dB from 7 to 9 GHz, and 11.5 dB from 9 to 11 GHz. Typical output power at 1-dB compression is +12.4 dBm from 5 to 7 GHz, +13.0 dBm from 7 to 9 GHz, and +13.0 dBm from 9 to 11 GHz. Typical output third-order intercept point (OIP3) is +25.0 dBm from 5 to 7 GHz, +24.5 dBm from 7 to 9 GHz, and +24.0 dBm from 9 to 11 GHz. The compact amplifier, fitted with female SMA connectors, features reliable unibody construction, measuring just 0.75 × 0.74 × 0.46 in. (18.80 × 19.05 × 11.68 mm) not including the lengths of the connectors. The LNA provides internal voltage regulation and reverse-voltage protection and boasts low current consumption, typically drawing 42 mA from a single 5-V dc supply. The 50-Ω amplifier is designed for operating temperatures from –40 to +85°C.

MINI-CIRCUITS, P.O. Box 350166, Brooklyn, NY 11235-0003; (718) 934-4500,
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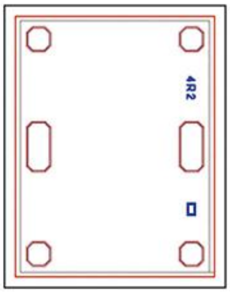

Tiny Splitter/Combiner Die Tackles 2 to 3 GHz

MODEL WP4R-D+ from Mini-Circuits is a broadband four-way, 0-deg. power splitter/combiner in die form targeting space-saving applications from 2 to 3 GHz. It provides 20-dB typical isolation between ports despite being only 0.805 × 1.032 mm in size. Manufactured with silicon MMIC integrated-passive-device (IPD) technology, the miniature four-way 0-deg. power splitter/combiner achieves impressive amplitude unbalance between signal paths of typically 0.1 dB, with typical phase unbalance of 2 deg. The full-band insertion loss is typically only 0.7 dB above the nominal 3-dB power split. Typical VSWR at all signal ports is 1.35:1. Well-suited for ISM-band applications, the RoHS-compliant, 50-Ω four-way power splitter/combiner die is designed for operating temperatures from –40 to +85°C. It handles as much as 1.5 W input power as a power splitter.

MINI-CIRCUITS, P.O. Box 350166, Brooklyn, NY 11235-0003; (718) 934-4500,
https://www.minicircuits.com/WebStore/dashboard.html?model=WP4R-DG%2B

Directional Coupler Spans 1 to 40 GHz

MODEL C-140-10 is a 10-dB stripline directional coupler with a wide frequency range of 1 to 40 GHz. It's targeted at test and broadband system applications. The coupler maintains 10-dB coupling with coupling flatness within ±1 dB. Maximum insertion loss is 2.5 dB, while the maximum input and output VSWR is 1.60:1. The directional coupler, which handles average power levels as high as 20 W and peak power levels to 3 kW, achieves minimum directivity of 10 dB, with average directivity of 20 dB across the frequency range. It's equipped with female 2.92-mm connectors and designed for operating temperatures from –32 to +85°C.

ELECTROMAGNETIC TECHNOLOGIES INDUSTRIES INC. (ET INDUSTRIES INC.), 50 Intervale Rd., Boonton, NJ, 07005, (973) 394-1719, FAX: (973) 394-1710, E-mail: sales@etiworld.com, www.etiworld.com

Compact Module Packs Dual Up/Downconverters

MODEL MC-6500 contains two independent frequency-upconverter/downcoverter sections capable of as much as 2-GHz real-time bandwidth from 1 to 18 GHz (and higher, with options). It's well-suited for radar, electronic-warfare (EW), and spectrum-analysis/monitoring applications. The downconversion channels translate signals from 1 to 18 GHz to L-band outputs at a 1.75-GHz center frequency (other L-band frequencies are available as options). The upconversion channels translate the same 1.75-GHz L-band signals to an output frequency range of 1 to 18 GHz. With a built-in antialiasing filter, unwanted signals are rejected by 80 dB. Spurious content is held to typically –60 dBc for both upconversion and downconversion. The frequency downconverters provide 20-dB gain with ±1-dB gain flatness, 15-dB noise figure, and 1-dB input compression point of –5 dBm. The frequency upconverters boast 20-dB gain with ±1.5-dB gain flatness, 15-dB noise figure, and 1-dB compression point of +15 dBm. The downconverters draw 9 W dc power per channel, while the upconverters require 11 W power per channel. The devices control group-delay variations to a mere 4 ns. Options include frequency coverage to 26.5 and 40.0 GHz. Standard units are supplied in a 6U VPX commercial-off-the-shelf (COTS) housing, with standalone, ruggedized hermetic enclosures available as an option. Matching dual frequency synthesizers, either PLL or faster DDS sources, in support of the frequency converters are also available as options.

FEI-ELCOM TECH INC., 260 Union St., Northvale, NJ 07647; (201) 767-8030, FAX: (201) 767-1326,
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(Continued from page 26)

BUILDING A RESILIENT NETWORK INFRASTRUCTURE

HUBER+SUHNER *(www.hubersuhner.com)* has been meeting the needs of mission-critical applications in some of the most rugged environments for over 15 years. The company understands that end-to-end RFoF systems are ultimately required to achieve the performance, stability, and reliability requirements of defense applications. This has led to the HUBER+SUHNER Polatis ultra-reliable all-optical switching solution, which delivers the specifications and features required to achieve maximum signal integrity throughout the system.

A CONNECTED BATTLEFIELD FOR THE FUTURE

The integration of C4ISR systems into existing platforms is driving RFoF system market growth on a massive scale, and this is set to continue well into the future. The military is looking to shift its focus from organizational and operational concepts to a military service that's fully integrated with connectivity systems to gather, process, display, and transmit information. This shift will create more opportunities for the military to use the capabilities of C4ISR applications. Critical to such performance is the network infrastructure that they choose to deploy. Using RFoF systems, the military can build a network that has greater resilience and performance, and one that's more secure at an overall lower cost.

HUBER+SUHNER connectivity solutions for the connected battlefield include both discrete and system-oriented solutions optimized for mission-critical applications. HUBER+SUHNER has worked on innovating the best-in-class RF and fiber-optic components and systems needed for end-to-end RFoF solutions. The HUBER+SUHNER portfolio of defense applications and solutions includes RF cables, connectors and RF-to-optical transceivers, multiplexers if running different channels on a single fiber, and a host of fiber and RF cable management products. mw

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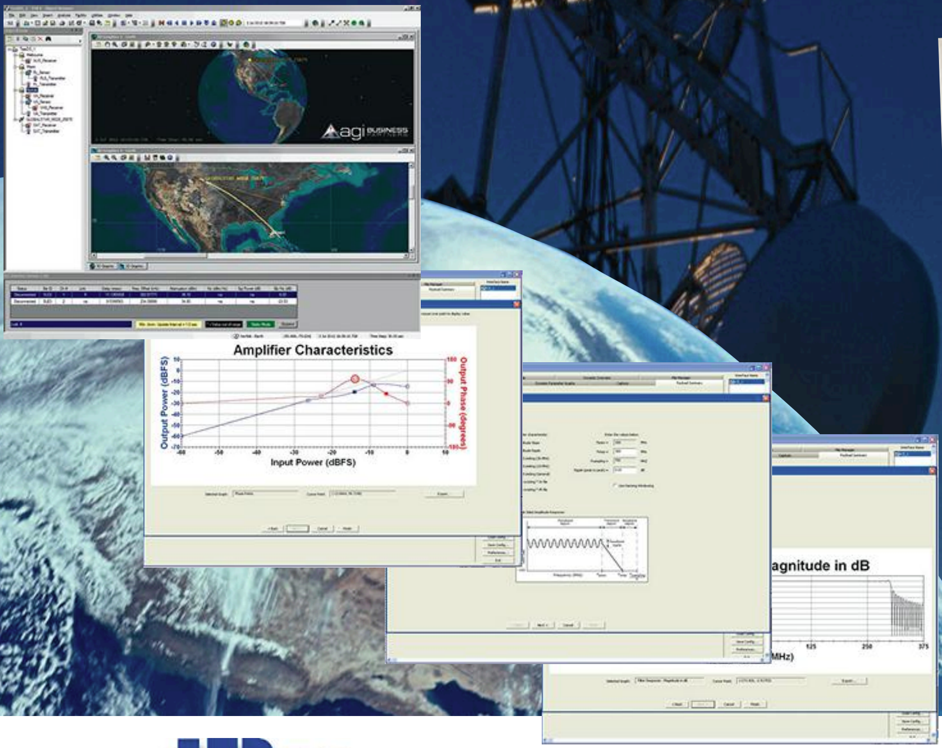

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