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Microwaves & RF

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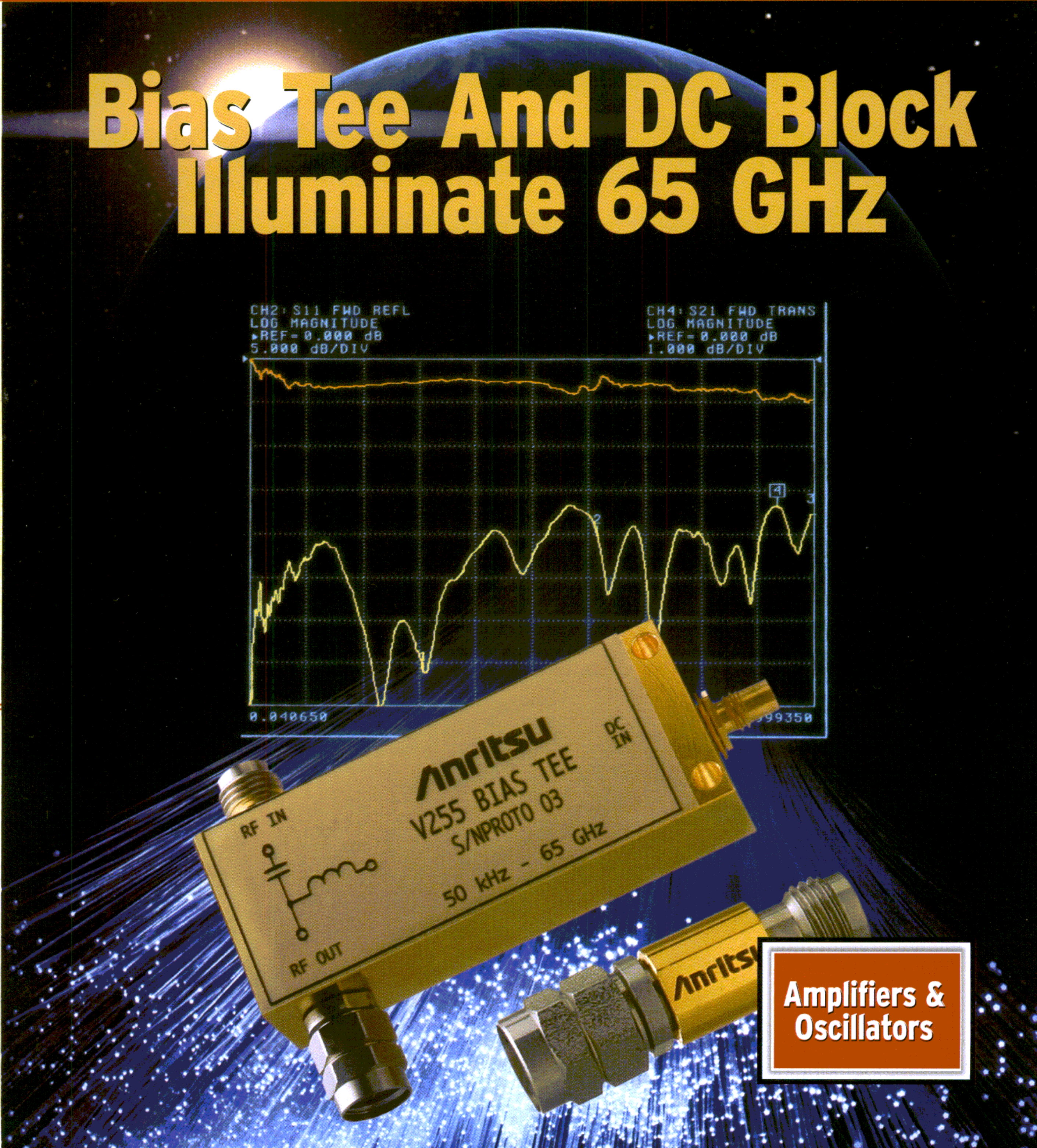
Facing Warfare In The Third Millennium

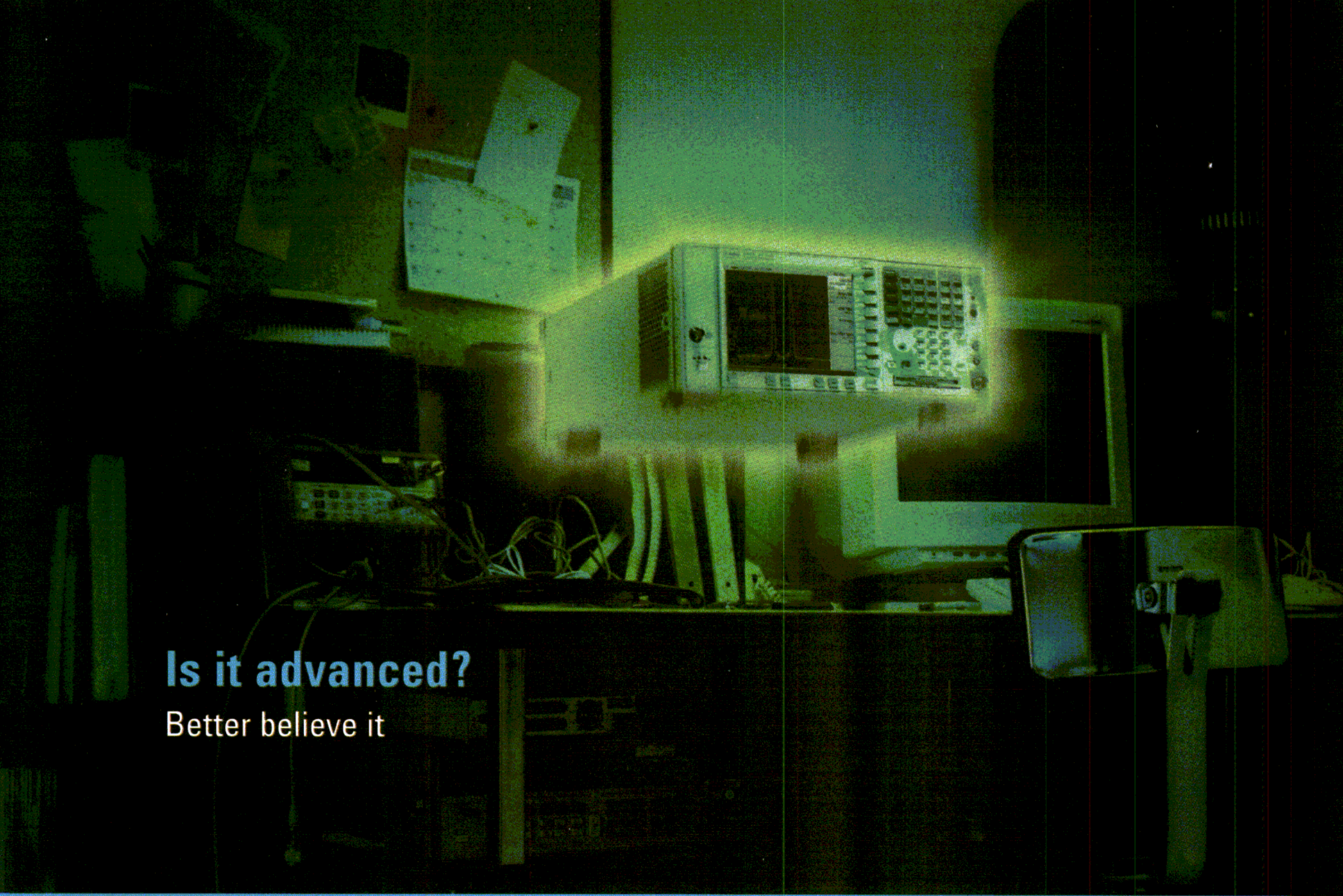
Design Feature
Simulation Tool Models And Verifies Jitter In Oscillators

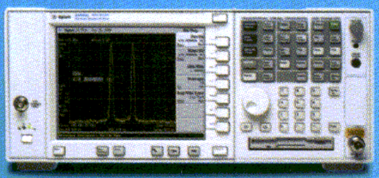
Product Technology
Linear HBT Amplifiers Arrive From New Source

Bias Tee And DC Block Illuminate 65 GHz



Amplifiers & Oscillators





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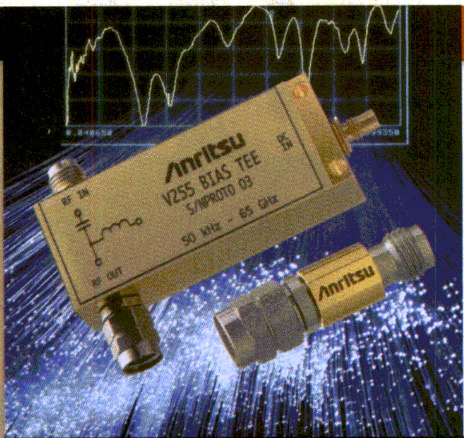

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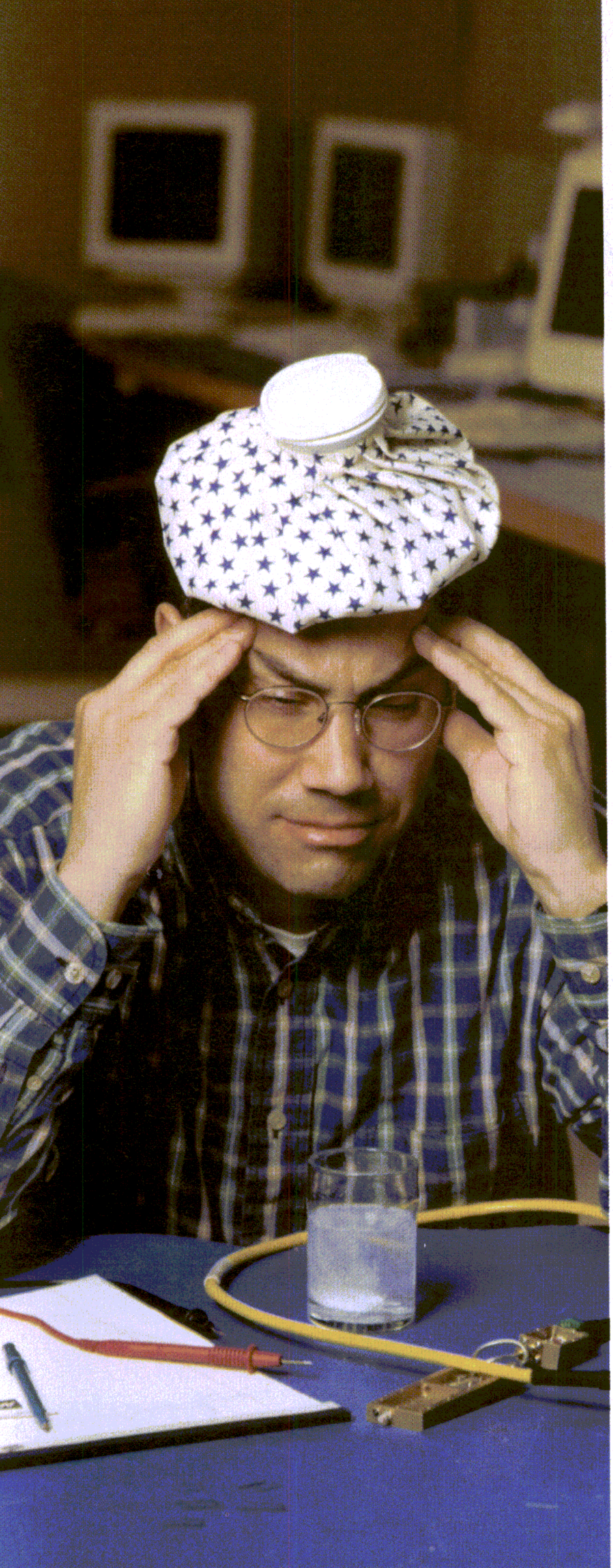

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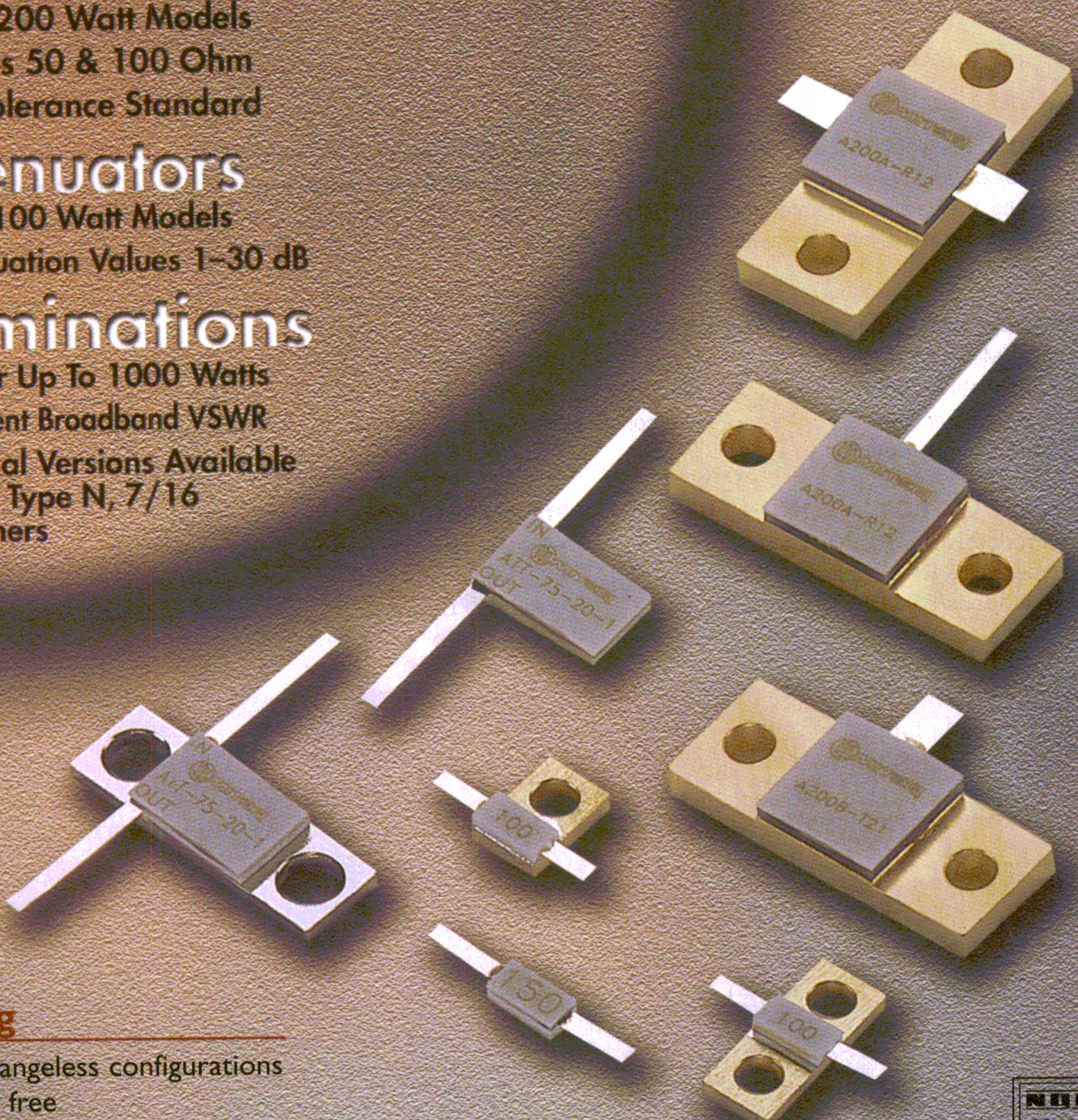

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Author Response

I WOULD LIKE to express thanks for the comments made by Phil Karn in last month's feedback on the article "Raise Bandwidth Efficiency With Sine-Wave-Modulation VMSK." (April, p. 79) My sincere thanks also to H.R. Walker on his comments in response to Phil Karn.

In the recent past, the urgent need for digital data transmission through existing analog channels has led to the development of various transmission techniques. Initially, these techniques, which included BPSK modulation, focused on moving digital information through the analog channel. Later, as demand grew for more transmission capacity, these techniques began to incorporate mechanisms for bandwidth reduction.

Two of the best known methods for bandwidth compression are QAM and multiple PSK. These are defined as NRZ line-code methods, since they concentrate the signal energy in a manner that enables them to be transmitted over wire lines. The earlier biphase technique is variable PSK. Another technique is the very MSK. This new version of VMSK modulation enables the earlier modulation biphase schemes to be combined to achieve bandwidth efficiencies that exceed 15 b/s/Hz. This is achieved at RF by using biphase encoding with the technique of SSB suppressed-carrier transmission.

All of the biphase codes are time varying. That is, they have a zero crossing point that varies with time. This waveform can be defined by a Fourier series that has a base frequency equal to the bit rate plus odd harmonics, along with a varying low-frequency amplitude component depending on the data pattern. They are polar codes and theoretically have few DC components or none at all. If a DC component is present, it can be disregarded in most cases.

One of the characteristics of NRZ line-code methods is that when the methods are transmitted with the use of carrier, they concentrate the spectrum of the transmitted signal around the carrier. In order to improve the spectral efficiency, only one of the two islands is required for transmission. This requires the use of a very sharp filter or, at least, a vestigal sideband filter that would impose serious implementation difficulties along with phase and amplitude distortion.

Editor's Note: K.H. Sayhood, the writer of this letter, is author of the April article "Raise Bandwidth Efficiency With Sine-Wave-Modulation VMSK" (p. 79). He is writing in response to Phil Karn's Feedback letter that appeared last month. Due to space constraints, the letter on this page will continue into next month's Feedback.

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Of Microwaves And The Military

Military influence has been long and strong on this industry, given its virtual birth during the first rumblings of World War II. Ever since the early days, when the Varian brothers were warming up those first microwave klystrons, microwave engineers have supported military electronic systems, generally for defensive applications. It is safe to say that the technological advances developed by the engineering talent in this country have provided us with extreme advantages in the many conflicts since World War II.

> *Troop morale needs bolstering, and as important, our military's technological edge requires some sharpening, having become dull during eight years of neglect.*

Microwave manufacturers have long sought that magic commercial application that would finally free them from the rigors of doing business with military customers. The microwave oven offered great promise in the 1970s and DBS television, with its highly visible 3-m antenna dishes, looked like that great commercial application in the 1980s. Wireless applications, of course, have proven to be the first truly far-reaching commercial application to benefit large numbers of microwave manufacturers. The sheer sizes of these markets (such as cellular networks and digital radios) have made military customers less appealing for many microwave suppliers, and many have turned their backs on military business. In most cases, this abandonment was probably for good reason, given the drastically decreased level of military funding during President Clinton's two terms of office.

The good news for the military, and for microwave suppliers still willing to conduct business with military customers, is the awareness on the part of the current administration and President George W. Bush of the need for increased military funding. Troop morale certainly needs bolstering, and as important, our military's technological edge requires some sharpening, having become dull and rusty during eight years of neglect.

On page 31, Fred Levien presents a strong case for the need of increased military funding. Professor Levien, founding Chairman of the Information Warfare Curriculum at the Naval Postgraduate School (Monterey, CA) and a retired US Navy Commander, is well-equipped to make an assessment of current military needs.

Levien's report is based on his Keynote Address at this year's first Military Electronics Show (Baltimore, MD, April 25, 2001). During that talk, he captivated his audience with the harsh reality of how much the military had been ignored during the past eight years. Hopefully, President Bush will be true to his campaign promises and restore our military to its rightful place as world leader.

Jack Browne

Publisher/Editor

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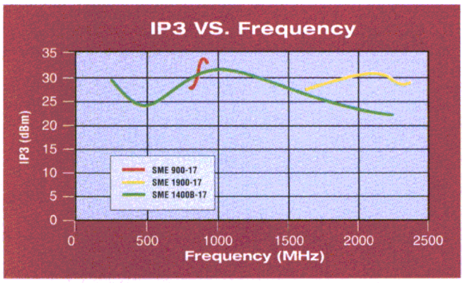

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SME 1400B-10	1-2200	1-2200	1-2000	+10	+6	+19	6.5	30
SME 1400B-13	1-2200	1-2200	1-2000	+13	+9	+22	6.5	30
SME 1400B-17	1-2200	1-2200	1-2000	+17	+13	+27	6.5	30
SME 1900-17	1600-2400	1400-2390	10-250	+17	+14	+29	7.4	26

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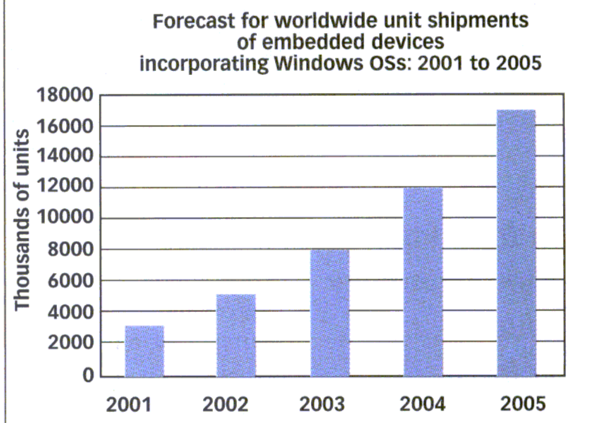
Windows Embedded Operating Systems Experience Growth

NATICK, MA—A study by Venture Development Corp. (VDC), a provider of market research to the embedded-systems industry, reveals that certain segments of the embedded market have been slow to adopt Windows Embedded operating systems. However, Microsoft is achieving acceptance in the embedded market with Windows CE in the consumer electronics segment.

VDC research shows that one of the main factors driving Microsoft's success in the consumer electronics market is the influence of the Microsoft brand. In addition, the popularity of the Windows platform in corporations' information-technology (IT) infrastructure and among computer users who are familiar with Windows desktop software is driving Windows CE acceptance in the handheld computer market.

One of the critical applications for Windows CE (Pocket PC) is the personal-digital-assistant (PDA) application. Pocket PC-based devices are expanding market share in this computing segment and VDC estimates that by 2005 they will account for 55 percent of all embedded devices shipped with Windows Embedded operating systems **(see figure)**.



The title of the VDC study is "Windows NT/NTE/CE 2000: Threats and Opportunities in the Embedded Systems Market." Findings are based on surveys mailed and e-mailed to 13,000 developers of embedded systems, interviews with engineers, product managers, and CTOs at leading software vendors and third-party service providers, as well as extensive secondary research.

Fiber-Optic Market To Drive Infrastructure Components To Billions In European Sales

OYSTER BAY, NY—As fiber-optic networks are built out in Europe, thanks to a new era of competition, those selling the components for the networks will realize billions of dollars in sales during the next five years, according to a series of reports from Allied Business Intelligence, Inc. (ABI).

"The European telecommunications market has always been considered a few years behind the US, but in the case of fiber optics this is not true, as the region is catching up quickly," says Mark Liggio, ABI's senior vice president of Broadband Communications. "With the more recent deregulation, the progress in the building of the European Union and a more conducive geography—the population centers are closely tied together—many carriers have seen the opportunity for using fiber optics to serve the pan-European market."

The European fiber-optics market is heavily advancing, with developments in components such as switches, transmitters (Txs), amplifiers, and receivers (Rxs). ABI's new report service, "The European Fiber Optic Deployment Series," examines the European market for eight fiber-optic components.

The series has found that by 2005, the total European market for semiconductor optical amplifiers will reach $138 million, the next-generation optical infrastructure market will reach $921 million, and the dense-wavelength-division-multiplexer (DWDM) market will reach $753 million.

Technology And Telecommunications Professionals Are Needed Despite Poor Economy

CLEVELAND, OH—According to the *Los Angeles Times*, a recent national poll of more than 1400 chief information officers forecasts a 21-percent increase in information-technology (IT) hires in the second quarter of this year, a rate of increase that is virtually unchanged from the previous two quarters. In Texas, for example, the need for technology talent is so strong that 34,000 jobs are unfilled, reports *The Dallas Morning News*.

"Technology is here to stay and it's going to grow and evolve," says Daniel Fager-George, manager of business development for BizSpace, Inc. "Companies will continue to look for talented professionals in all areas of telecommunications and technology because each day brings a new discovery."

"The economy may be a little slow right now, but it's not dead," says Michael Forrest, president of JobOptions.com. "In the long run, the technology revolution is going to drive the economy and transform our lives."

"In the long run, the technology revolution is going to drive the economy and transform our lives."

Active Microwave Modules Market In North America To Exceed $3.4 Billion In 2005

HUNTINGTON, YORK, UNITED KINGDOM—A report in a new series from Engalco, "Microwaves North America III—Active Microwave Modules—Markets to 2005," forecasts that the overall total available markets for this class of microwave products will grow from the expected $2.4 billion level this year to exceed $3.4 billion in 2005. The study includes detailed market data on the following classes of microwave products: electronic switches [chiefly monolithic microwave integrated circuit (MMIC) based], voltage-controlled oscillators (VCOs), dielectric-resonator oscillators (DROs), yttrium-iron-garnet (YIG) oscillators, linear amplifiers, log amplifiers, wireless local-area-network (WLAN) chip sets, frequency synthesizers, and other relatively complex function models. There are full profiles of 13 major players, as well as an extensive industry directory.

Every year, a substantial [illegible] quency synthesizers with m[illegible] types of modules forecast to be w[illegible] $750 million this year and to exc[illegible] in 2005. Next ranked are WLAN c[illegible] and linear amplifiers, in that order.

In most instances, the current economic slowdown is having only marginal effects on these markets that are mainly fueled by the strong growth in sectors such as broadband satellite, 3G mobile, and WLANs. There are also substantial and slowly extending opportunities in the defense sector.

Processors Support Several Industry-Standard Wireless Protocols

PLEASANTON, CA—embedded wireless devices, inc. has unveiled the e8024™ Voice and Data Broadband Wireless Gateway Processor and the e9024™ Voice and Data Wireless LAN processor. The e8024 and e9024 processors are the first devices capable of concurrently supporting the IEEE 802.11a, operating at 5.7 GHz, as well as industry-standard wireless protocols such as IEEE 802.11b, Bluetooth™, HomeRF, and HiperLAN2.

Designed for residential and enterprise broadband wireless gateways and wireless local-area networks (WLANs), these products eliminate Quality of Service (QoS), interference, and performance problems that occur when the IEEE 802.11b protocol, with its direct sequencing scheme, co-exists on a wireless network with the other 5-GHz spectrum and 2.4-GHz industry standards. The e8024 and e9024 processors provide additional bandwidth and multiple basebands to support simultaneous data and voice at the broadband and premise-side distribution interfaces.

Next-generation systems will use the 5-GHz spectrum to address these interference and performance issues. By allocating a broad 200-MHz band (versus 85 MHz for 802.11b), the 5-GHz spectrum enables higher data throughput and eliminates interference from 2.4-GHz-based appliances such as microwave ovens and cordless telephones. Similar to the 2.4-GHz spectrum, 5-GHz spectrum does not require a license throughout much of the world.

Report Details Impacts Of Competition On Local-Exchange Outside Plant Assets

AUSTIN, TX—Technology Futures, Inc. (TFI), a telecommunications forecasting firm, has published a report entitled, *The Impacts of Competition and Technology on Local Exchange Outside Plant Assets*. The report was authored by TFI's president, Lawrence K. Vanston, Ph.D. Vanston's report quantifies the impact of broadband technology and wireless competition on the ILECs' copper (Cu) cable plant.

"DSL technology is allowing the ILECs to compete with cable companies in providing residential broadband access. In the short run, this increases the utilization of some copper pairs, but the overall impact of broadband is to reduce the amount of copper used," remarks Dr. Vanston.

"DSL technology is allowing the ILECs to compete with cable companies in providing residential broadband access."

Dr. Vanston continues, "By including forecasts of all the important factors, we get a comprehensive picture of what is really happening to copper in the ILEC outside plant in the long term. This is the only way you can realistically estimate the impact of all these changes on the value of ILEC investment."

The report also quantifies the substitution of wireless for wire-line voice service, forecasting the displacement of ILEC access lines and usage. In addition, it assesses potential competition for cable voice and the Internet, including voice-over-Internet protocol (VoIP).

There were several key findings in the study. By year-end 2004, more than 25 percent of US households will have adopted broadband services, up from approximately 5 percent at the end of 2000. By 2010, that percentage will likely exceed 60 percent. Due to competition from wireless, cable modems, and digital-subscriber-line (DSL) services, the number of ILEC-provisioned narrowband access lines will begin to decline sharply around 2003. When competitive and technology impacts are considered, existing metallic cable has an average remaining life of between 5.2 and 9.0 years, with 7.0 years most likely. And finally, the need for ILECs to provide broadband services with data rates above 1.5 Mb/s will drive the adoption of fiber-in-the-loop. Forecasts have indicated that 1.5 Mb/s will meet the typical user's expectations until about 2005, the 6 Mb/s will do so until around 2010, then 24 Mb/s until 2015, and finally 100 Mb/s thereafter.

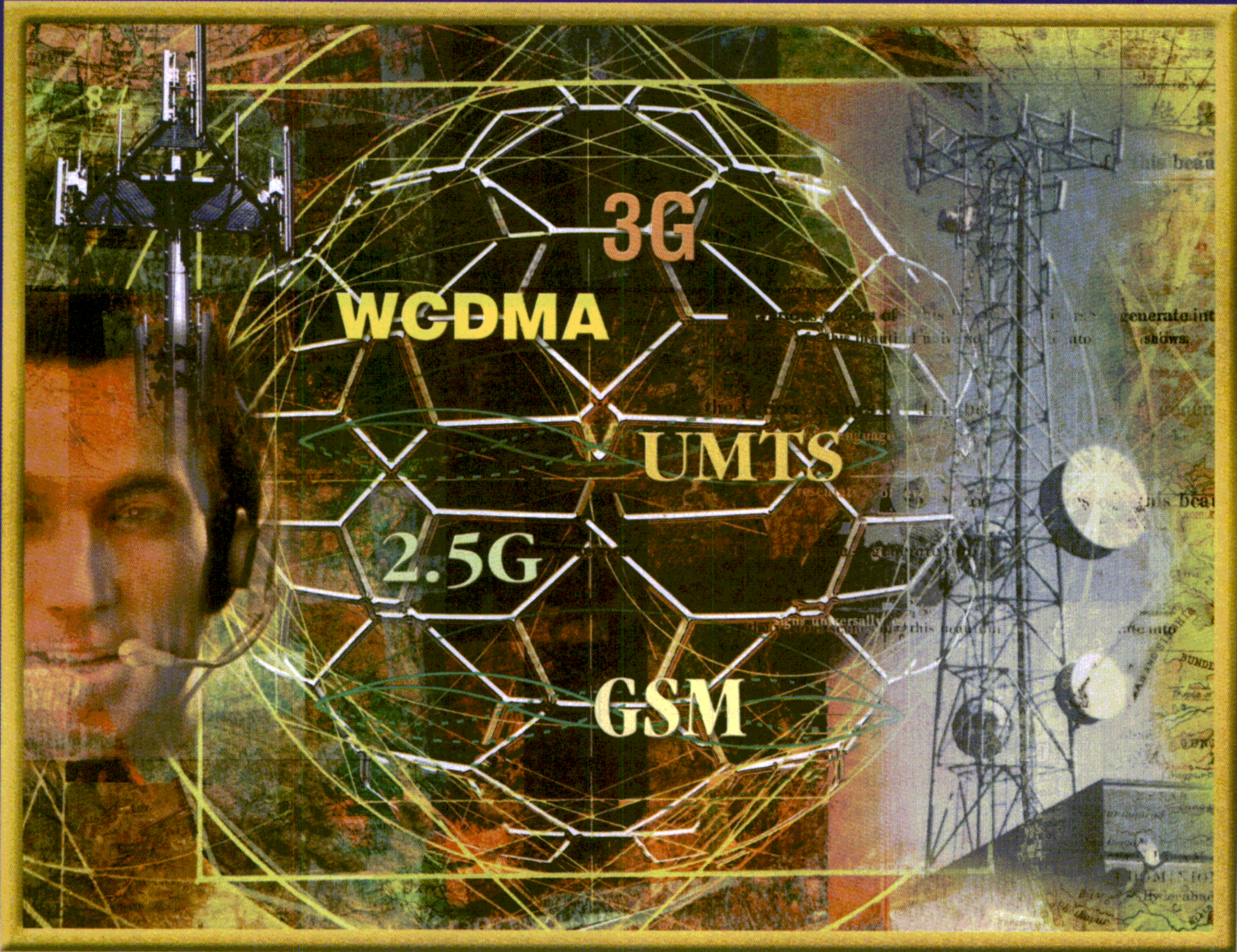
RF Exposure Levels To Workers Are Being Studied

HAPPAUGE, NY—With occupational safety a legal and health concern for employers all over the world, the ability to document dangerous and disputable incidents can only serve to clarify the facts of a case. In a litigious society such as ours, every piece of information counts.

This is especially true in areas of occupational safety that are still poorly understood, such as RF safety. With the explosion of wireless communications, satellite broadcasting, and other forms of RF-emitting devices, the possibilities of worker exposure to dangerous levels of RF radiation levels are greater than ever. While many people are aware of the RF hazards facing broadcast and telecommunications workers, exposure to excessive RF radiation levels, as established by government agencies, is also possible in many industrial processes that use microwave and other RF heating devices. These include furniture, semiconductor, auto and aircraft parts manufacturing, as well as bookbinding, lumber drying, ceramics, coal, and many other raw- and finished-goods production. Building workers and others with the opportunity to work near or around radio-transmission equipment may also risk exposure to non-compliant levels of RF energy. Building workers and others with the opportunity to work near or around radio-transmission equipment may also risk exposure to non-compliant levels of RF energy.

"Even residential buildings these days rent out rooftop space to wireless carriers, especially in urban areas," says Richard Strickland, director of business development for Narda Safety Test Solutions, which manufactures RF safety devices. "So now even building-maintenance workers who never dreamed of RF exposure as one of their job hazards are at risk these days."

Narda has taken a step toward enabling employers to log crucial information about RF exposure levels to workers. In a first for the industry, its new line of Nardalert XT personal RF monitors enables users to create and store a continuous data log of RF exposure levels. MRF



Linearity. Reliability. Power!

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NEWS

Facing Warfare In The Third Millennium

Growing potential threats from developing nations and a declining investment in US military forces have led to a situation where the US and allies are vulnerable to hostile actions.

FRED LEVIEN
President
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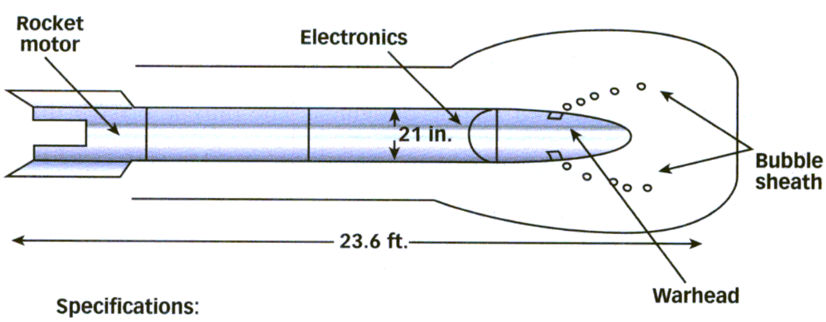
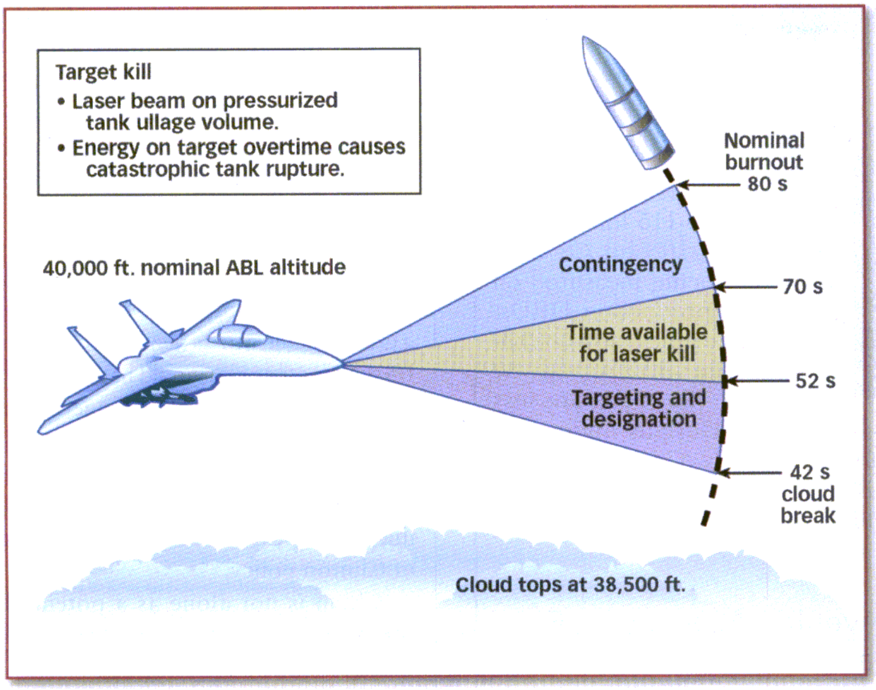
Military solutions continue to grow more advanced in the third millennium. Radical changes within this country and the ever-changing balance of world political alliances have combined to create an environment where modern military solutions must be agile and flexible. However, by examining the past, it may be possible to anticipate the military needs to come.

Political environments change constantly. With the "fall" of the Soviet Union, one great threat to this country has been "minimized." Still, another threat is in the making. One of the key changes in the world's political environment over the last several years is China's role on world peace. A panel of international experts recently pointed out the potential for danger from the Chinese political situation: "...eight years [1992-2000] of strategic obtuseness and a policy of appeasement has emboldened China to such a significant degree that the next few years could well prove to be an extreme-



Specifications:
Total weight = 4845 lbs.
Warhead weight = 91 lbs. (nuclear capable)
Range ≈ 11 to 19 miles
Propulsion = originally: 1350-kW hydrogen-peroxine turbine
*Modified: high-speed underwater ("SKVAL") rocket model APR-2
*New speed = 230 mph
Entered service = Year 2000
Developer: Morteplotekhnika Research Institute

1. The Russian Skval submarine torpedo can achieve an underwater speed of 230 mph.



2. The Air Borne Laser (ABL) weapon system has been developed by Boeing, TRW, and Lockheed Martin under contract to the US Air Force.

ly dangerous period in US/China relations."[1] In addition, the panel found "a China much more dangerous now than it ever was when he [former President Bill Clinton] arrived in 1993."[1]

The last decade was one where international relationships have slipped but, at the same time, the US military has declined. The decade witnessed a severe drop in US troop morale and, while a simple examination of numbers does not tell a full story, it does reveal the extent of US military erosion resulting from the financial fallout of the previous presidential administration's budgets **(Tables 1 to 3)**. Even during a decade of financial bounty, the Clinton Administration chose to apportion the smallest percentage of US Gross Domestic Product to the defense of the nation since prior to World War II **(Table 4)**. The net result on the US military forces was pervasive. The US Air Force decreased from 36 to 10 fighter wings. The Navy dropped from

Table 1: A shrinking US Navy

	1990	1999	CHANGE
Budget (in constant 2000 dollars)	$146 billion	$84 billion	–42 percent
Personnel (total)	1,322,500	882,200	–33 percent
Navy	582,900	372,300	
US Marine Corp active	196,700	172,200	
Navy reserve	149,900	90,800	
US Marine Corp reserve	44,500	40,000	
Civilian	349,000	206,900	
Strategic submarines (ballistic missile)	34	18	–47 percent
Tactical submarines (attack)	91	57	–37 percent
Aircraft carriers	14	12	–14 percent
Major surface combatants	206	116	–44 percent
Navy fighter and attack aircraft	759	468	–38 percent
US Marine Corp fighter and attack aircraft	450	328	–27 percent

approximately 600 ships to slightly more than 300. Approximately three-quarters of a million troops were dropped from the active US forces. US Army divisions alone have decreased from 18 to 10, making the US Army now only the seventh largest in the world.

Ross Munro writes that, "The Clinton legacy in Asia has been to weaken America's standing and to make China a greater danger to its neighbors and the US than it otherwise would have been."[1] His conclusion is that the danger from China as a military threat is not just potential and long term in nature. It is "here and now, real and present...constituting one of the most serious indictments of Clinton policy."[1]

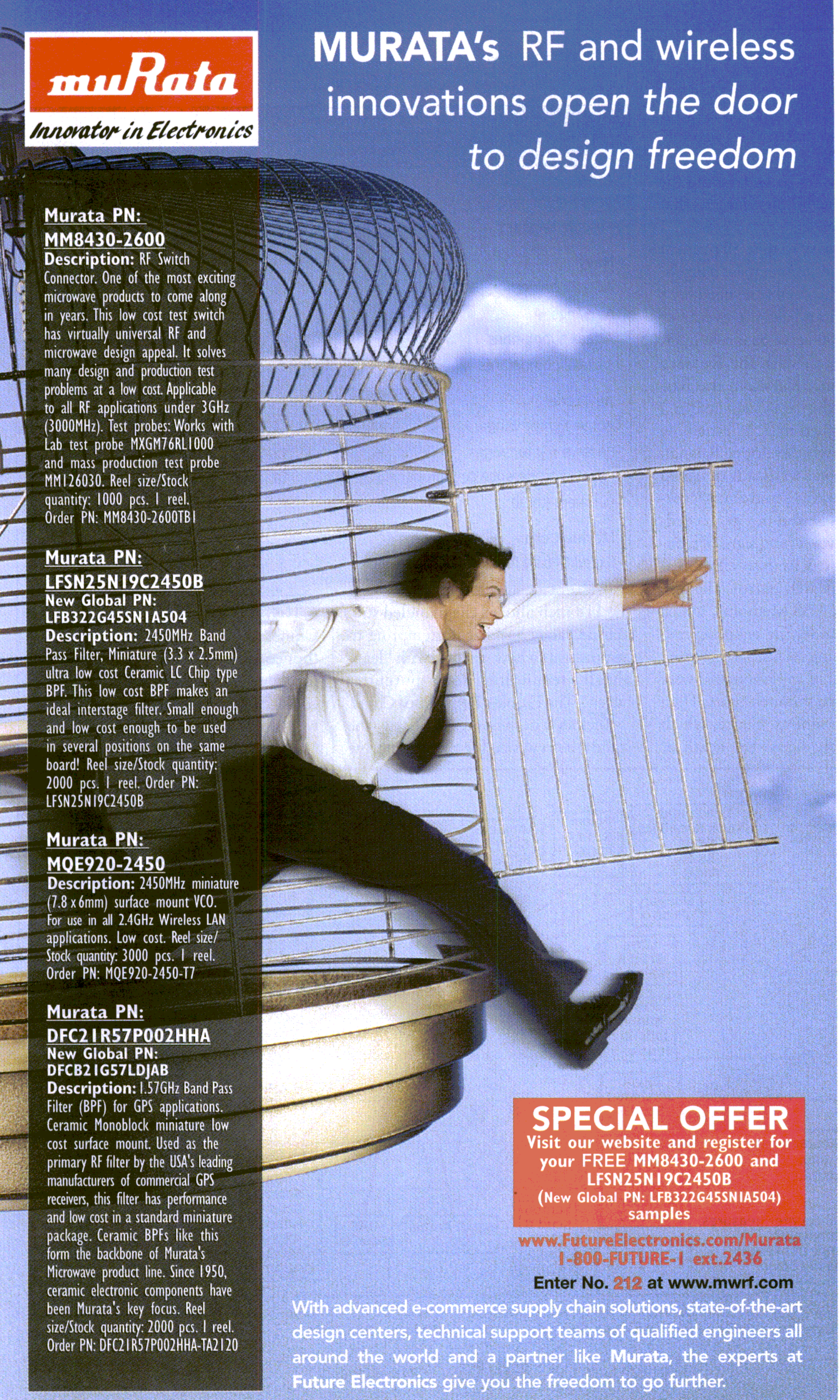
China is not alone as a potential threat. Growing world powers (and

Fortunately, There is a sense under the current administration that much greater attention will be paid to the US military.

not to discount Russia) have become unfriendly to the West, and to the US in particular. These political changes occurred at a time when the US armed forces were becoming less capable with each succeeding Department of Defense (DoD) budget. This is the legacy that the recent eight years of leadership have passed on to the current administration in particular and to the country at large.

Fortunately, there is a sense under the current administration that much greater attention will be paid to the US military. Unfortunately, reversing this decline in military strength and the equally troublesome underlying damage to troop morale will not come easily or inexpensively.

These next three points need to be emphasized: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Murata PN: DFC21R57P002HHA
New Global PN: DFCB21G57LDJAB
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1. There is an absolute necessity for the US to develop strategic "leap-ahead" advances in military technology.

2. It is necessary to understand the possibility of the military dangers that this nation faces.

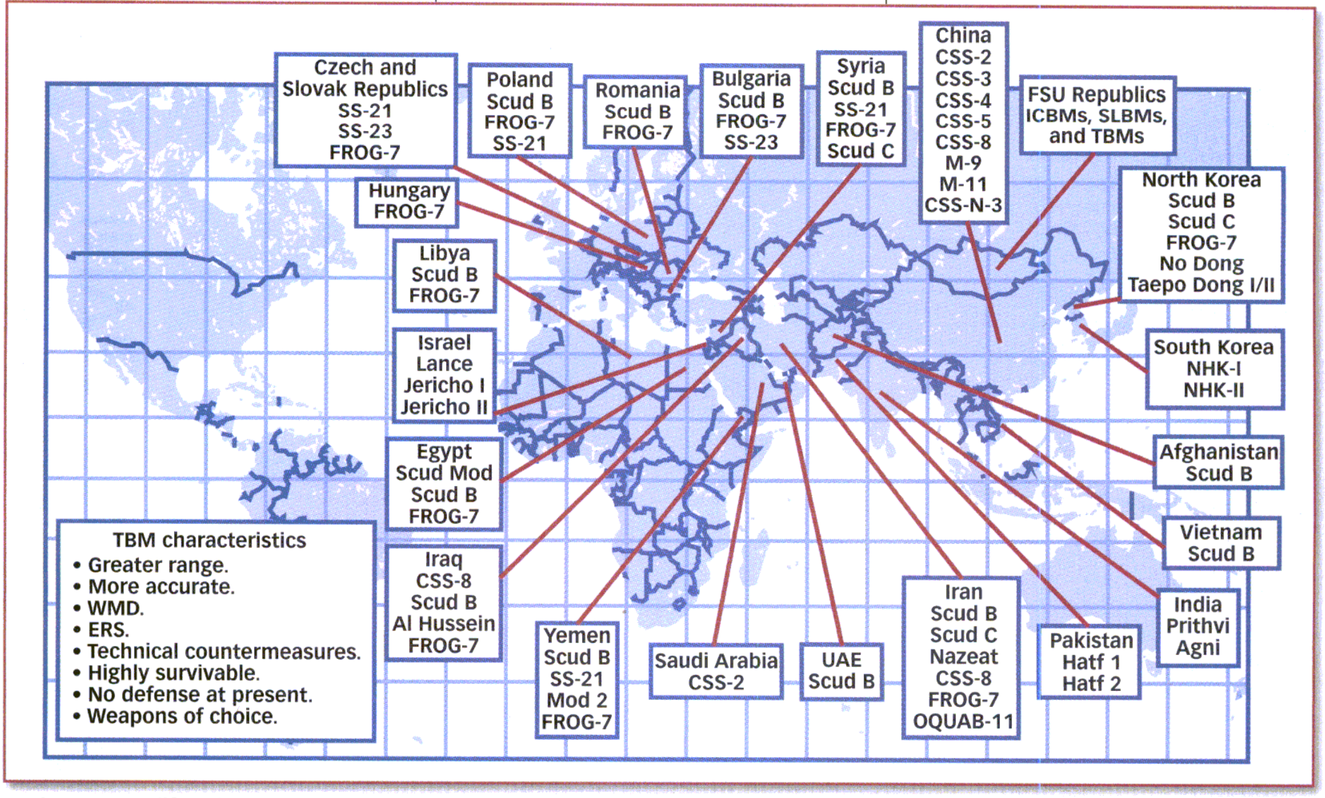
3. It is necessary to take all prudent steps necessary to counter these dangers.

For those who doubt the need for a strong US military, let the events of August 12, 2000 serve as a reminder. That story began with the Russian Naval submarine, Kursk, in the Barents Sea. The ship was engaged in a war game designed to determine how best to use a newly developed Russian undersea weapon specifically designed to sink US submarines and aircraft carriers. The exercise did not go according to plan, however, and signs of trouble were detected by two US Naval submarines in nearby patrol. The USS Toledo and the USS Memphis, which were close by under the sea observing the exercise using MASINT listening gear, heard a pair of explosions that were detected just two minutes apart. Then there was an ominous silence, which would inevitably signal the death of the 118 men aboard the mangled submarine. They had been attempting to demonstrate a new weapon with "leap-ahead" technology for a potential buyer—The People's Republic of China.

Included in the oversized crew (the standard crew size for this submarine is 108) it is believed there had been a Chinese Naval officer observing the (hoped for) performance of this new torpedo. As an unclassified guess, it could be pieced together that the new torpedo, known as "Skval" in Russian ("Squall" in English), failed, allowing

Table 2: Main elements of the US Armed Forces

	1990	2000
Army		
Active divisions	18	10
Reserve brigades	57	42
Navy		
Aircraft carriers	15	11
Air wings	13	10
Attack submarines	91	55
Surface combatants	206	116
Air Force		
Active fighter wings	24	12
Reserve fighter wings	12	7
Reserve air-defense squadrons	14	4
Bombers (total)	277	190
Marine Corp		
Expeditionary forces	3	3



3. The TBM proliferation worldwide includes more accurate weapons with increased range.

fuel to leak into the submarine. The fuel ignited a fire in the forward torpedo room, which caused the torpedo to explode. This led to a more extensive fire, which a few minutes later detonated all the other ordnance aboard the Kursk located in the forward torpedo room. These explosions ultimately resulted in the sinking of the vessel.

This scenario was essentially confirmed by a note found on the body of a Kursk sailor recovered from the wreckage. Not surprisingly, Russian naval officers, despite all of the evidence to the contrary, have continued to insist that a collision with a foreign submarine was the first in the series of events that doomed the Kursk.

In the details of this new weapon **(Fig. 1)**, there is one number that dominates—the underwater speed. The torpedo, which was driven by a jet-rocket engine, is reported to have had an estimated speed of 230 mph. This tremendous speed is made possible by generating an envelope of bubbles forward of the torpedo nose, thereby reducing the frictional drag through the water. This is five times the underwater speed of any known torpedo currently in any Navy's inventory. It is believed that a more advanced version of the Skval torpedo is under development, with an underwater speed of 300 mph. The impact of this potent nuclear-capable weapon was best described by a Jane's Defense Week analyst who said that "Skval is a weapon that has the prospects of destroying entire Naval battle groups at once. This could abruptly blow a hole in US Carrier-based air superiority. The consequences are grave. This new torpedo has the potential to tilt the balance of power."[3]

***Despite** the harsh economic situation, Russia has employed some of the best engineers and scientists, and is still capable of developing "leap-ahead" technologies.*

The potential customer for this technology is from the same nation that recently took out a US Navy EP-3E aircraft over international waters. As the crew tried to keep it from crashing, China arrogantly blamed the US for intruding into Chinese air space. China has demanded and is still demanding an apology for their aggressive, illegal, and hostile actions. This is a nation that held the US in high regard as a poten-

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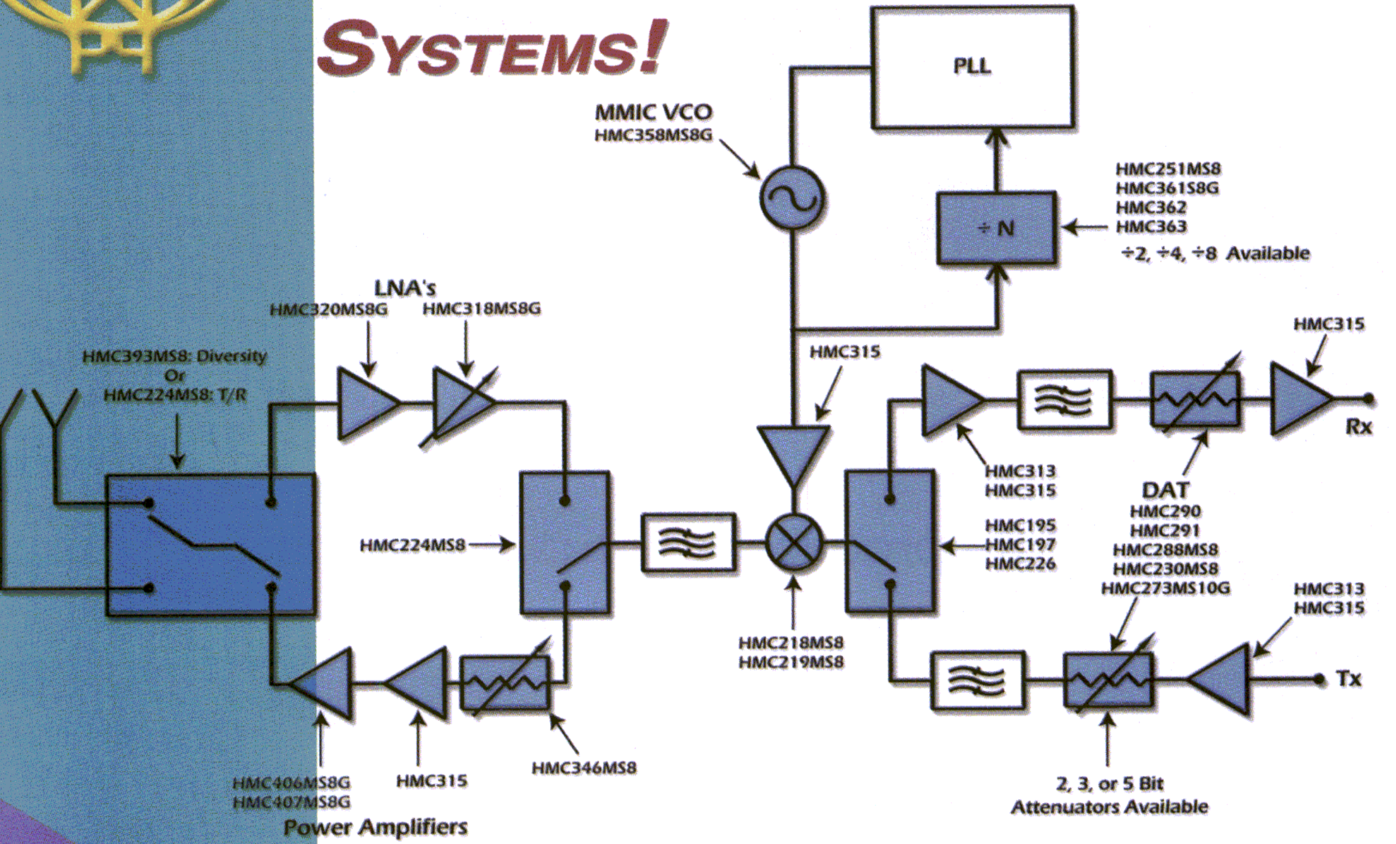

Frequency	Product Type	Part Number
DC - 6.0 GHz	LNA	HMC313
DC - 7.0 GHz	LNA	HMC315
5.0 - 6.0 GHz	LNA	HMC318MS8G
5.0 - 6.0 GHz	LNA	HMC320MS8G
5.0 - 7.0 GHz	Driver Amplifier	HMC407MS8G
5.0 - 6.0 GHz	Power Amplifier	HMC406MS8G
3.0 - 6.5 GHz	Divider	HMC251MS8
DC - 10.0 GHz	Divider	HMC361S8G
DC - 11.0 GHz	Divider	HMC362
DC - 12.0 GHz	Divider	HMC363
5.6 - 6.2 GHz	VCO	HMC358MS8G
4.5 - 6.0 GHz	Mixer	HMC218MS8
4.5 - 9.0 GHz	Mixer	HMC219MS8
0.7 - 3.7 GHz	Digital Attenuator	HMC288MS8
0.7 - 3.7 GHz	Digital Attenuator	HMC273MS10G
0.7 - 4.0 GHz	Digital Attenuator	HMC290
0.7 - 4.0 GHz	Digital Attenuator	HMC291
0.75 - 2.0 GHz	Digital Attenuator	HMC230MS8
DC - 8.0 GHz	VVA	HMC346MS8G
5.0 - 6.0 GHz	Diversity Switch	HMC393MS8G
DC - 3.0 GHz	SPDT Switch	HMC197
DC - 2.0 GHz	T/R Switch	HMC226
DC - 2.5 GHz	T/R Switch	HMC195
5.0 - 6.0 GHz	T/R Switch	HMC224MS8

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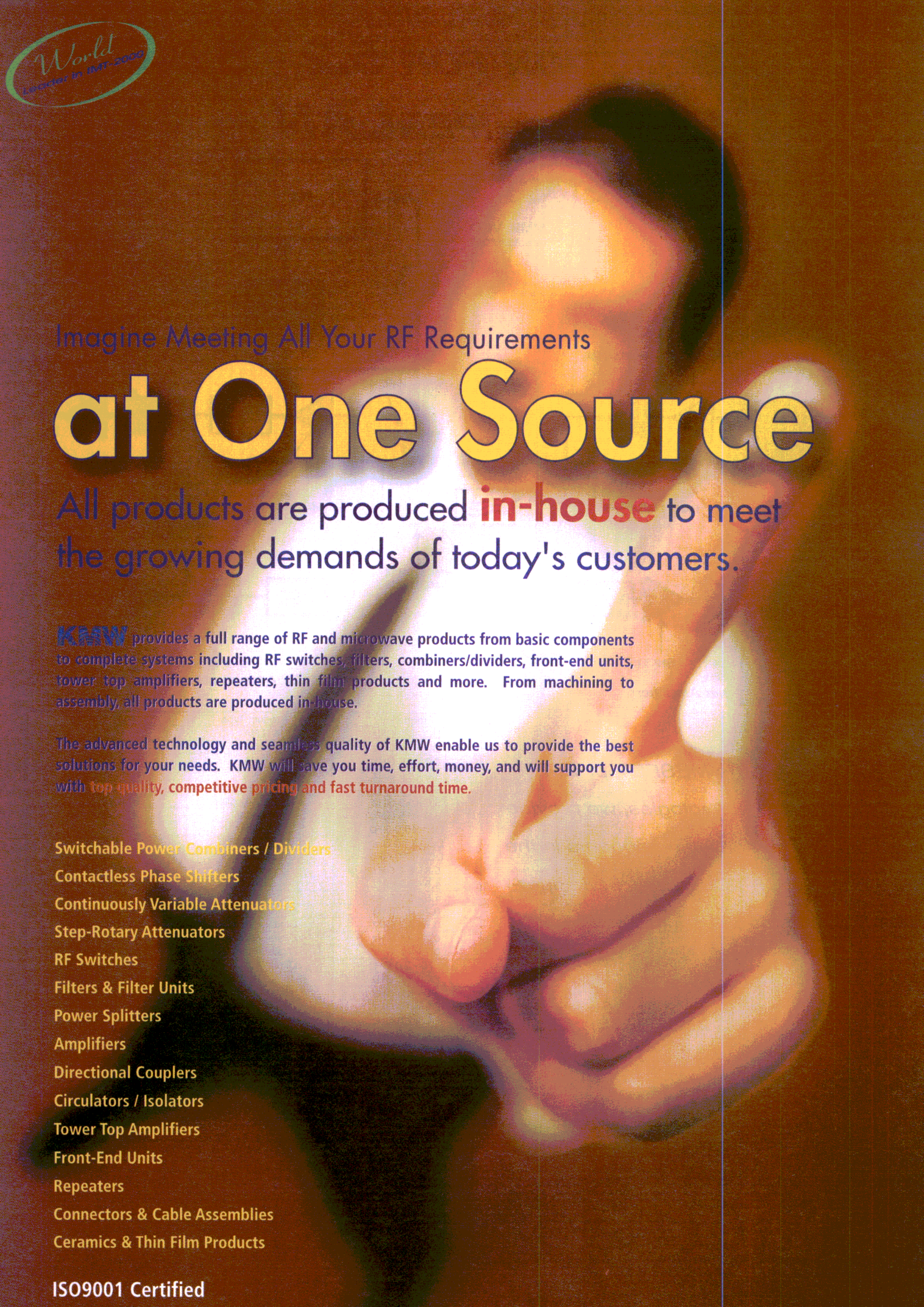
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tial friend only eight years ago.

The Kursk incident highlights how two major communist nations have allied together to form a potential threat. The technologically advanced Russians are in need of a "customer," while the Chinese have shown with the EP-3E incident that they are willing to steal whatever technology they cannot buy. Despite the harsh economic situation, Russia has employed some of the world's best engineers and scientists, and is still capable of developing "leap-ahead" technologies. And in its decline, Russian has shown that it will sell anything to anyone at any time.

Table 3: A shrinking Air Force

	1990	1999	CHANGE
Budget (in constant 2000 dollars)	$116 billion	$78 billion	–33 percent
Personnel (total)	992,300	715,000	–28 percent
Active	539,300	365,900	
Reserve	197,600	181,200	
Civilian	255,400	168,700	
Heavy bombers	244	143	–41 percent
Fighter and attack aircraft (total)	2610	1455	–44 percent
Active	1743	906	
Reserve	867	549	
Airlift aircraft (total)	851	756	–9 percent
Intertheater	401	331	
Intratheater	450	425	

The existence of the Skval torpedo is a sign that the US does not own all of the most advanced technologies. With the growing competition for weapons technology, the rising importance of Asia becomes obvious. The Skval represents "asymmetric warfare." It is essentially a $200,000 torpedo that can obliterate a US battle group that costs multiple billions of dollars, and causes fatalities to more than 6000 US service men and women.

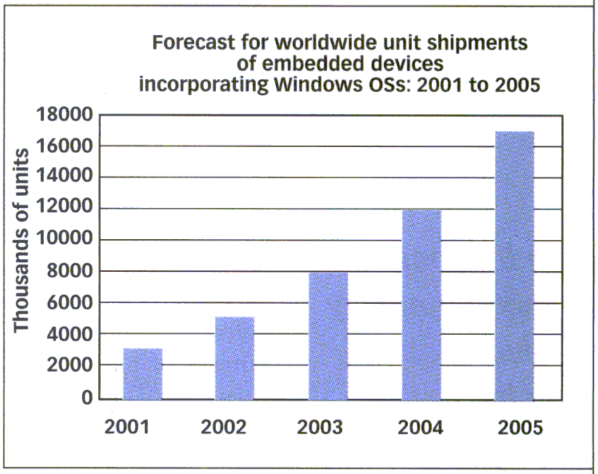
The Kursk story effectively supports a report prepared by Andy Marshall for President Bush. The report, entitled *A Strategy for a Long Peace* and subtitled *Quick Look*, examines what lies ahead for the US military. Marshall, who has served as a mentor to the author during the development of his Information Warfare (IW) course for the Naval Postgraduate School, has worked at the Pentagon for more than 50 years. He was chosen for the study due to his highly regarded technical expertise and legendary skills in big-

Windows Embedded Operating Systems Experience Growth

NATICK, MA—A study by Venture Development Corp. (VDC), a provider of market research to the embedded-systems industry, reveals that certain segments of the embedded market have been slow to adopt Windows Embedded operating systems. However, Microsoft is achieving acceptance in the embedded market with Windows CE in the consumer electronics segment.

VDC research shows that one of the main factors driving Microsoft's success in the consumer electronics market is the influence of the Microsoft brand. In addition, the popularity of the Windows platform in corporations' information-technology (IT) infrastructure and among computer users who are familiar with Windows desktop software is driving Windows CE acceptance in the handheld computer market.

One of the critical applications for Windows CE (Pocket PC) is the personal-digital-assistant (PDA) application. Pocket PC-based devices are expanding market share in this computing segment and VDC estimates that by 2005 they will account for 55 percent of all embedded devices shipped with Windows Embedded operating systems **(see figure)**.



The title of the VDC study is "Windows NT/NTE/CE 2000: Threats and Opportunities in the Embedded Systems Market." Findings are based on surveys mailed and e-mailed to 13,000 developers of embedded systems, interviews with engineers, product managers, and CTOs at leading software vendors and third-party service providers, as well as extensive secondary research.

Fiber-Optic Market To Drive Infrastructure Components To Billions In European Sales

OYSTER BAY, NY—As fiber-optic networks are built out in Europe, thanks to a new era of competition, those selling the components for the networks will realize billions of dollars in sales during the next five years, according to a series of reports from Allied Business Intelligence, Inc. (ABI).

"The European telecommunications market has always been considered a few years behind the US, but in the case of fiber optics this is not true, as the region is catching up quickly," says Mark Liggio, ABI's senior vice president of Broadband Communications. "With the more recent deregulation, the progress in the building of the European Union and a more conducive geography—the population centers are closely tied together—many carriers have seen the opportunity for using fiber optics to serve the pan-European market."

The European fiber-optics market is heavily advancing, with developments in components such as switches, transmitters (Txs), amplifiers, and receivers (Rxs). ABI's new report service, "The European Fiber Optic Deployment Series," examines the European market for eight fiber-optic components.

The series has found that by 2005, the total European market for semiconductor optical amplifiers will reach \$138 million, the next-generation optical infrastructure market will reach \$921 million, and the dense-wavelength-division-multiplexer (DWDM) market will reach \$753 million.

Technology And Telecommunications Professionals Are Needed Despite Poor Economy

CLEVELAND, OH—According to the *Los Angeles Times,* a recent national poll of more than 1400 chief information officers forecasts a 21-percent increase in information-technology (IT) hires in the second quarter of this year, a rate of increase that is virtually unchanged from the previous two quarters. In Texas, for example, the need for technology talent is so strong that 34,000 jobs are unfilled, reports *The Dallas Morning News.*

"Technology is here to stay and it's going to grow and evolve," says Daniel Fager-George, manager of business development for Biz-Space, Inc. "Companies will continue to look for talented professionals in all areas of telecommunications and technology because each day brings a new discovery."

"The economy may be a little slow right now, but it's not dead," says Michael Forrest, president of JobOptions.com. "In the long run, the technology revolution is going to drive the economy and transform our lives."

"In the long run, the technology revolution is going to drive the economy and transform our lives."

Active Microwave Modules Market In North America To Exceed $3.4 Billion In 2005

HUNTINGTON, YORK, UNITED KINGDOM—A report in a new series from Engalco, "Microwaves North America III—Active Microwave Modules—Markets to 2005," forecasts that the overall total available markets for this class of microwave products will grow from the expected $2.4 billion level this year to exceed $3.4 billion in 2005. The study includes detailed market data on the following classes of microwave products: electronic switches [chiefly monolithic microwave integrated circuit (MMIC) based], voltage-controlled oscillators (VCOs), dielectric-resonator oscillators (DROs), yttrium-iron-garnet (YIG) oscillators, linear amplifiers, log amplifiers, wireless local-area-network (WLAN) chip sets, frequency synthesizers, and other relatively complex function models. There are full profiles of 13 major players, as well as an extensive industry directory.

Every year, a substantial lead is taken by frequency synthesizers with markets for these types of modules forecast to be worth more than $750 million this year and to exceed $1 billion in 2005. Next ranked are WLAN chip sets and linear amplifiers, in that order.

In most instances, the current economic slowdown is having only marginal effects on these markets that are mainly fueled by the strong growth in sectors such as broadband satellite, 3G mobile, and WLANs. There are also substantial and slowly extending opportunities in the defense sector.

Processors Support Several Industry-Standard Wireless Protocols

PLEASANTON, CA—embedded wireless devices, inc. has unveiled the e8024™ Voice and Data Broadband Wireless Gateway Processor and the e9024™ Voice and Data Wireless LAN processor. The e8024 and e9024 processors are the first devices capable of concurrently supporting the IEEE 802.11a, operating at 5.7 GHz, as well as industry-standard wireless protocols such as IEEE 802.11b, Bluetooth™, HomeRF, and HiperLAN2.

Designed for residential and enterprise broadband wireless gateways and wireless local-area networks (WLANs), these products eliminate Quality of Service (QoS), interference, and performance problems that occur when the IEEE 802.11b protocol, with its direct sequencing scheme, co-exists on a wireless network with the other 5-GHz spectrum and 2.4-GHz industry standards. The e8024 and e9024 processors provide additional bandwidth and multiple basebands to support simultaneous data and voice at the broadband and premise-side distribution interfaces.

Next-generation systems will use the 5-GHz spectrum to address these interference and performance issues. By allocating a broad 200-MHz band (versus 85 MHz for 802.11b), the 5-GHz spectrum enables higher data throughput and eliminates interference from 2.4-GHz-based appliances such as microwave ovens and cordless telephones. Similar to the 2.4-GHz spectrum, 5-GHz spectrum does not require a license throughout much of the world.

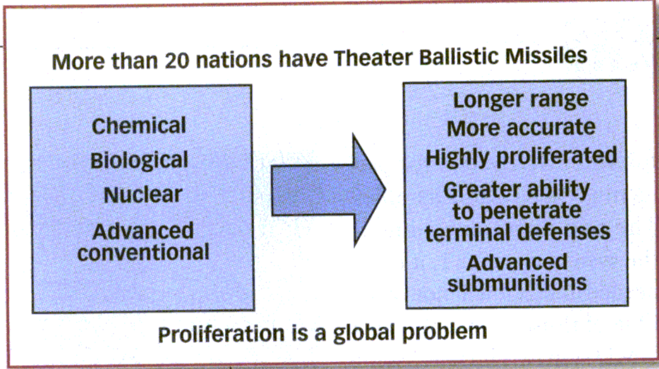
picture strategic thinking.

In reviewing the 25-page first draft of the report, it is possible to target key areas that are deemed to be crucial to understanding and planning for the future of the US Military. The report notes the trends in political powers, keying on rising competition in Asia. This translates into the high-priority need for defending the US homeland with a National Missile Defense (NMD) system. The report also highlights the increased risk incurred by forward basing of US military forces on land and naval forces close offshore. The report reviews primary tools, including IW, and the impact on the US Armed Services. The report calls for support of the Naval submarine service and the development of a new breed of small "street fighter"

For Those who would dispute the need for an NMD system, the pages of US newspapers have provided clues as far back as 1997 as to the need.

vessels, while de-emphasizing Navy aircraft carriers, even suggesting a cut in numbers. The report urges US Air Force dominance in the electromagnetic (EM) spectrum, suggesting that this should be the primary mission of the B-52 fleet and the B-1 bombers. The report opposes the call for the newly proposed JSF fighter A/C, with unmanned air-combat vehicles suggested as a replacement for US Air Force support missions. In the realm of promising warfare technologies, the report makes note of advanced radar satellites, undersea power projection, and new IW technologies such as directed-energy weapons and computer-network warfare.

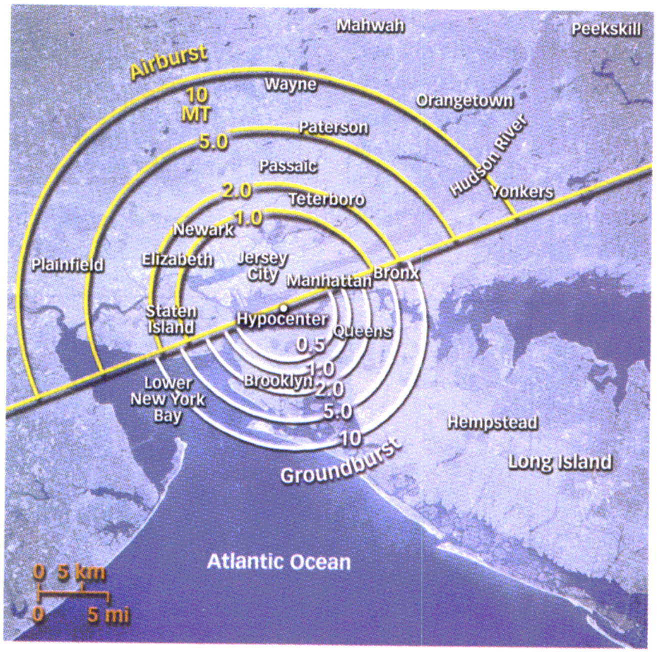
One more potential weapon system is covered by President Bush's plans



4. The threat of TBM capabilities worldwide extends to a variety of payloads, including nuclear, biological, and chemical weapons.

for an NMD system for protecting the US and its allies. The two main objections to the system are that 1. it is not needed and 2. it will not work. One of those objecting to the NMD system is French Assembly President Paul Quiles, who denounced NMD as a "military program unworthy of the name." He adds that "this system will never work...the technology is not there."[4]

The growth of potential threats worldwide would seem to dismiss the first objection. The technological history of the US would certainly dispel the second. Those who doubt that the system is possible should be reminded of all those who doubted that the US could land a man on the moon.



5. The numbers next to the circles denote the explosive yield in megatons, and the circle itself covers a metropolitan area of New York City.

The US has already come close to the technological reality of this anti-missile system. By applying a deliberate, carefully planned, and superbly managed effort, the US Air Force has maintained an on-time program that is within budget and meets all key technical milestones. With less than 40 s to acquire, track, and kill an Intercontinental Ballistic Missile (ICBM) target in the boost phase **(Fig. 2)**, the anti-ballistic-laser (ABL) margin for error is extremely small. Yet, to date, all tests point to a successful program completion in a relatively short span of time. Fullup system design has resulted in the first flyable airframe that is being assembled now, with a deployable system possible within three years.

For those who would dispute the need for an NMD system, the pages of US newspapers have provided clues as far back as 1997 as to the need. One newspaper title in particular, cautioned "Ballistic Missiles within easy reach for many Nations."[5] A recent Air Force overview of missile capability presently deployed around the world provides an accurate and sobering view of more than 10,000 ballistic missiles (TBMs) currently in place. The status shown for these missiles is by type of missile and the country in possession of it **(Fig. 3)**. Not all of the nations listed are friendly toward the US and the payloads that these countries are prepared to deliver are deadly and constantly being improved **(Fig. 4)**.

A graphic depiction is presented here of the potential damage from a missile strike **(Fig. 5)**.[6] The effects of detonating a single nuclear weapon over the southernmost tip of Manhattan island in New York City, a spot known as Battery Park, are portrayed. Two choices are provided to calculate the effects of a single armed nuclear-missile attack on New York City: a surface detonation and an air burst (as might be expected from an incoming ICBM). For example, in using a one-megaton value for an air burst caused by a present-day multiple independently targeted re-entry vehicle (MIRV), most of the population within the depicted circle would be killed by thermal radiant exposure. A single half-megaton ground blast would result in approximately three to four million fatalities.

Readers should be aware of the real and present dangers in our unsettled world. They are urged to apply their talents to the goal of developing advanced "leap-ahead" concepts and technology in support of next-generation military electronics. This can lead to better ways of providing the tools that the US military desperately needs to ensure our country's survival.

Editor's Note: This article is an excerpt of a Keynote Address presented by the author at the First Military Electronics Show in Baltimore, MD, April 24, 2001. MRF

Table 4: Federal spending by administration (percent of gross domestic product)

PRESIDENT (FISCAL YEARS)	SOCIAL WELFARE	NET INTEREST	DEFENSE	OTHER	TOTAL
Nixon (1970–1975)	9.4	1.4	6.7	2.6	20.1
Ford (1976–1977)	11.9	1.6	5.2	3.0	21.7
Carter (1978–1981)	11.6	1.9	5.0	3.3	21.8
Reagan (1982–1985)	12.2	2.9	6.2	2.5	23.8
Reagan (1986–1989)	11.2	3.2	6.2	2.0	22.6
Bush (1990–1993)	12.6	3.4	5.1	2.3	23.4
Clinton (1994–1997)	14.0	3.4	2.8	1.6	22.8

Source: Congressional Research Service.

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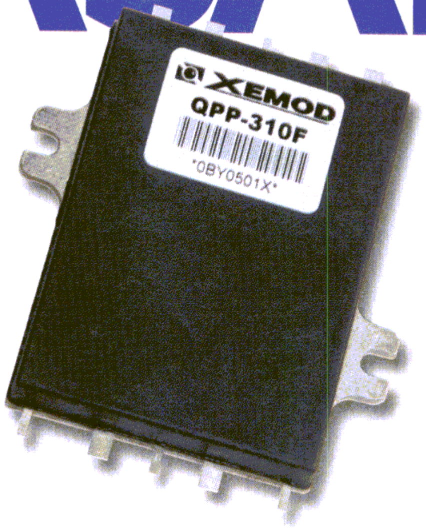

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Dr. Zoltan Cendes is Chairman and Chief Technology Officer of Ansoft Corp. (Pittsburgh, PA), and is also an Adjunct Professor of Electrical and Computer Engineering at Carnegie Mellon University (Pittsburgh, PA). He is a member of the IEEE Microwave Theory & Techniques (MTT) Technical Committee on computer-aided engineering (CAD) and is on the editorial board of the International Journal of RF and Microwave Computer-Aided Engineering. He has served on the International Steering Committee of the COMPUMAG Conference, and is a past chairman of the IEEE Conference on Electromagnetic Field Computation.

An Interview with Dr. Zoltan Cendes

MRF: What led you to start Ansoft?

Cendes: I started Ansoft in 1984 while I was a professor at Carnegie Mellon University conducting research. Some of our research partners were asking us to produce commercial software tools in addition to the continuing research they traditionally funded us to perform, so we began to satisfy their requests.

MRF: What was Ansoft's first product?

Cendes: We called it Eddy. It was a two-dimensional program for computing eddy currents in AC conductors. We followed that effort with some simple two-dimensional microwave simulation programs. Soon thereafter, Hewlett-Packard Co. asked Ansoft to develop a three-dimensional finite-element simulator that they would sell, having been asked by their customers for alternatives to test-and-measurement instrumentation to speed the product-development process. We were obviously excited about the opportunity, and it contributed heavily to our expansion into higher-frequency design.

At that time, solving high-frequency electromagnetic problems using finite elements often resulted in unphysical spurious modes. This was a severe limitation, and presented us with a great challenge. We invented procedures to avoid these problems, which made it possible to solve three-dimensional electromagnetic field problems using finite elements for the first time.

MRF: What was the result of the work for HP?

Cendes: The culmination of our work was the software program HFSS, which was the first commercial program that could simulate complex three-dimensional geometries. After we finished the initial software development, HP marketed the product exclusively for us. We began shipping the product in 1990, and it has since become the industry standard for three-dimensional, electromagnetic field simulation.

HFSS continues to be one of our core products, but of course it's much faster now—15,000 times faster than in 1990. With greater speed, an ability to address open-region problems and create full-wave SPICE models, HFSS is utilized by designers simulating complex antenna structures, optical-communications systems, RF ICs, high-speed PCBs and connectors, and biomedical applications. Perhaps the most significant change is in optimization and sensitivity analysis, where you can now synthesize improvements by specifying design goals and simultaneously investigating manufacturing changes to reduce costs.

MRF: The Ansoft/HP (now Agilent Technologies) partnership came to an end in the mid 1990s. What happened?

Cendes: We parted ways with HP about six years ago, basically because our views of HFSS development differed. However, we both continued to develop HFSS products, and now Agilent has agreed that our HFSS technology is best, and they have sold their HFSS business to us. We are working with them to transition their customers to Ansoft solutions. HP (now Agilent) was key to the early development of our company, and we're happy we'll be working together again.

MRF: What is the significance of your Full-Wave Simulation Program with Integrated-Circuit Emphasis (SPICE) product?

Cendes: Full-Wave SPICE is a unique functionality that addresses one of the key areas that high-speed digital designers have struggled with during the past several years. As frequencies have approached and gone into the realm of RF, digital designers face new problems. SPICE, the main tool for circuit designers, does a poor job of handling high-frequency effects. However, using Ansoft's Full-Wave SPICE, digital designers can model the true physics in their transient SPICE circuit simulation—without resorting to either inaccurate single-frequency models or trial and error. Full-Wave SPICE automates the entire process and produces frequency-dependent SPICE models in Hspice, Pspice, or Maxwell SPICE formats, so that designers can continue to work with tools they are familiar with, and obtain the accuracy they need to address today's speeds.

MRF: What has been the obstacle to solving this problem?

Cendes: Traditional circuit models miss much of the physics and are best used for low-frequency circuits and for components that are small compared to wavelength. Designs for higher frequencies must account for electromagnetic interactions. If you solely rely on circuit-level models, the first iteration of a design will most likely not deliver the desired performance. This means you have to create a lengthy cycle of prototypes to achieve your goals.

MRF: You have described Ansoft Designer, which debuts next year, as one of the major benchmarks in your company's history. Why is this so?

Cendes: Ansoft Designer is a benchmark for Ansoft and a major step forward for the industry. It represents the next generation of design tools that integrate different levels of simulation technology, allowing the user to customize the level of accuracy needed at all phases of the design process. Our "solver on demand" architecture will greatly simplify the optimization process, making it far easier for engineers to make tough design decisions early in the process, and avoid surprises downstream.

For example, if you are solving a complicated circuit using components for which there is no model, Ansoft Designer will call the electromagnetic simulator on demand to perform the required simulation. Until now, people needed to rely on multiple tools and troublesome translators, and go back and forth trying to analyze the design. Designer lets you work all levels, simultaneously, without burdening the user with additional steps. This approach simplifies the process without constraining a designer's creativity, and the modern user interface makes it easy to learn.

MRF: What do you feel are the broad trends in design tools?

Cendes: The Holy Grail has always been simulation software that is reliable, highly accurate, easy to use, and robust so a broad spectrum of problems can be solved. However, it is a mistake for EDA vendors to believe they can make something very simple and yet highly versatile. There are always trade-offs. Our customers tell us that they want high accuracy—it's why they buy simulation in the first place. Next, they need the power to customize and tune our tools to their environment.

Finally, and just as important, no solution is complete without knowledgeable, responsive people to help designers apply these tools. It's often our applications people that see the next innovation, from their experience with customers using the tool. Service is often the key differentiator and it's the area that many simulation companies completely overlook or gravely understaff. I am proud of our commitment to service. I frequently receive accolades from our customers when I visit the MTT and DAC conference.

MRF: What are the challenges you face in achieving this goal?

Cendes: There are two parallel paths creating the ultimate solution. The first is the algorithm itself—every year we invent better algorithms that will give more accurate answers faster and can handle larger problems. The next level is the user interface, which we continually improve with customer feedback. Within the software level, many improvements are made in quality that the user never sees, but experiences indirectly in terms of superior performance, fewer inconsistencies, and better operation. I remember writing in FORTRAN, and how constrained that was. With the object-oriented programming that we have been using since 1988, we can create software structures that accommodate more sophisticated programming and better code more efficiently. Both of these paths are evolving every year, and they will continue to inch closer to an ideal solution.

MRF: How much is your development work driven by, or perhaps limited by, computer hardware?

Cendes: In our early days with HFSS, around 1990, our abilities really were dictated by the performance of the computer hardware. However, at about that time, workstation performance really began to accelerate without an increase in cost. That let us finally solve three-dimensional problems without incurring high additional costs.

As an example, one of our original test problems with HFSS was to develop a waveguide-to-coax adapter. This simple structure took 16 hours to solve on a Sun 4 workstation. Today we can solve the same problem in 20 seconds. Of course, some of that speed increase is attributable to hardware performance, but we continuously increase the performance of the software as well. On average, we have more than doubled the speed of our algorithms every year.

MRF: Are there problems that still require a workstation to solve?

Cendes: It's hard to say. Every time we reach a new level regarding the problem size we can solve, customers ask for more. Although PC performance has narrowed the gap, and can solve a great many problems, today's workstations have 64-B operating systems that allow almost limitless amounts of memory.

However, I believe customers will continue to push the limit, and there will always be a need for varying levels of computer performance. I'm quite certain that we haven't yet seen all the innovations of our time, and that hardware vendors will work hard to keep a differentiation that is valuable to the marketplace.

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It's nice that one doesn't have to rely on fancy prose when describing a measurement device. A few numbers will suffice instead. Take the FSU, for example – our new highend spectrum analyzer. Whether it's displayed average noise level, phase noise or large-signal characteristics: the FSU is at the top of the range in all disciplines of RF performance. Likewise with level accuracy. Naturally, if you offer such solutions, you are not permitted to only go half way. Therefore, we have enhanced the record performance with a suitable package of extras. For example: a complete line of detectors, a large collection of bandwidth resolutions and filter characteristics implemented for the first time to the upper limit of 50 MHz, tools for fast time-domain analysis etc, etc. And why all this? So that you, the developer or manufacturer of the next generation of communications technology, will have substantial reserves for tackling the most demanding tasks, or maximizing production yield. You will find that every dB more and every ms less are worthy benefits which will quickly pay for themselves. The specifications, by the way – you can find them on the Internet, of course. And the FSU is available immediately from your Rohde & Schwarz sales partner.

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MRF: Are there limitations to current high-frequency design software? Are there problems these tools still cannot solve?

Cendes: There are still significant challenges for our industry. The one that generates the most attention is tackling larger and larger RF ICs with thousands of transistors. Standard harmonic balance engines cannot handle very large structures, and while you can solve some in the time domain, the results are not very accurate. Traditional circuit-design tools for standard ICs address the larger structures, but not the microwave characteristics. Microwave tools cannot handle the level of complexity, as can standard IC design tools. We're working in from both ends of this technology to solve this problem.

Another challenge is in the realm of electromagnetic and circuit simulation. They can be done in cosimulation, but the electromagnetic part is still only working on a smaller portion of a device or circuit. If you simulate a whole device or board, it can be overwhelming for pure electromagnetic simulators. Our goal is to analyze the whole board with electromagnetic accuracy. I already see progress in handling much larger nonlinear circuits with harmonic balance, and in quickly doing much larger boards with many devices.

The technology incorporated into the Ensemble planar simulator, SVD Fast Solve, has allowed us to speed up full-wave simulation with planar structures by an order of magnitude.

MRF: Do you have any final thoughts on the direction of the high-frequency electronic-design-automation (EDA) industry?

Cendes: This is a very exciting time in high-frequency design automation. I started Ansoft with the vision to leverage the power of electromagnetics across component-, circuit-, and system-level design. Today, we are the recognized world leader in this area. By providing total solutions with circuit- and system-level models derived from accurate electromagnetic models, our tools replace long and expensive build-and-test cycles with fast and economical computer simulations. As more designs move to ever-higher frequencies, to millimeter-waves and to optics, the need for our solutions exponentially increase. Traditional methods no longer work well. Customer hunger for greater speed, greater density, longer battery life, and improved functionality is driving virtually every technology company, regardless of market. Ansoft is qualified to serve these needs, with our expert personnel and our pioneering work in creating products to serve this market. We have only just begun. I find the new world of high-frequency design very exciting, and can hardly wait to see the revolution in high-speed digital, RF, microwave, and optical products that will result. MRF

InGaP HBT AMPLIFIERS

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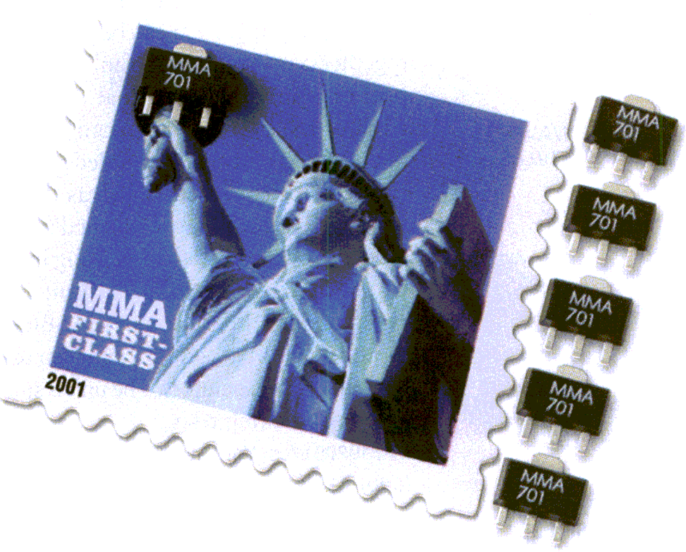

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HBT AMPLIFIERS

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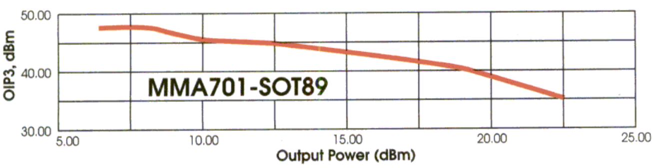
MMA701-SOT89	0.001–4.0	7.0	130	12.5	+27	+48	80
MMA710-SOT89	0.001–5.0	7.0	95	12.5	+22	+37	130

* Matched SSG at 2.0 GHz
** Measured at 2.0 GHz



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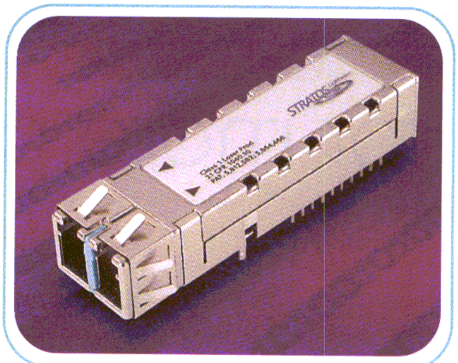

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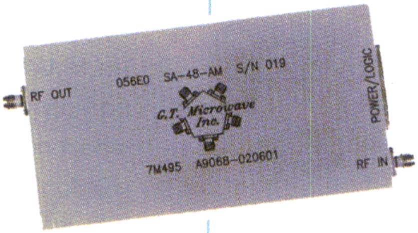

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3G Wireless Inches Forward

DESPITE FACING MAJOR financial obstacles, US telecommunications companies are ready to roll out early versions of 3G wireless communications in the next few months. Verizon Wireless and Sprint PCS, two of the nation's largest wireless service providers, plan to launch systems based on the Qualcomm cdma2000 technology later this year. Also this year, AT&T and Cingular Wireless intend to release a 2.5G service with features similar to cdma2000. These introductions come in the wake of predictions that universal 3G will be delayed a few years due to economic and technical setbacks affecting the worldwide communications industry.

One of the reasons behind the push to roll out cdma2000 in the US could be the looming battle between the two CDMA technologies that are fighting to control the world market: cdma2000 and WCDMA. Backers of each technology claim that theirs is superior and should be adopted as the universal standard. WCDMA appears to have the early lead based on support from Finland's Nokia, the leading manufacturer of wireless phones, and the major European telecommunications carriers. But the European telecommunications industry is burdened with debt, holding up its 3G deployment and opening a window of opportunity for cdma2000 to gain some adherents to its technology.

Already, more than 12 phone companies in North America, Asia, and South America with a subscriber base of 100 million customers have elected cdma2000. If these companies can bring their networks online quickly, it will provide a big lift to the market penetration of cdma2000.

Meanwhile, the technical battle between the two CDMA formats is still being waged. Key to the cdma2000 argument is that it makes more efficient use of the spectrum over which radio signals travel. WCDMA proponents are counting on the fact that the large installed GSM base in Europe and Asia can be converted into WCDMA technology. It may all boil down to spectrum, where cdma2000 has the edge since it can use its present airwaves for 3G, whereas WCDMA needs more spectrum than it presently owns. MRF

CONTRACTS

ANADIGICS—Has received production orders for CATV tuner ICs from Samsung for use in their digital set-top box. Samsung's DVB-RC set-top box will use ANADIGICS' ACU2109 upconverter IC, and ACD02202 downconverter dual-synthesizer IC, to provide a double-conversion RF tuner solution.

Harris Corp.—Announced that it, along with its consortium partners, has been awarded the first phase of a four-phase turnkey rebuild of TransAlta Utilties' microwave network. TransAlta is Canada's largest nonregulated electric generation company. The Harris-led consortium will engineer, furnish, install, commission, and cut-over TransAlta's current analog and digital microwave system to a completely digital broadband network using Harris' Constellation™ family of point-to-point radios. The total project, spanning the province of Alberta, is estimated to be worth $15 to $20 million over the next four years, and will include digital microwave, multiplexers, switches, DC power, tower infrastructure, and other civil works.

Raytheon Company's RF Components Division (RRFC)—Will supply LG Electronics Co. Ltd. with the RMPA 1951 GaAs HBT PA for the handset market. LG Electronics Co. Ltd., a Korean wireless telephone manufacturer, has earmarked the TM910 dual-band trimode handset for the US CDMA market.

LCC International, Inc.—Has been selected by China's Unicom Horizon Mobile Communications Corp. Ltd. to provide design services for their nationwide wireless network. The initial contract, valued at approximately $600,000, calls for LCC to assist Unicom in developing the design guidelines for their system, as well as perform the actual planning, RF engineering, and optimization services in Foshan City and Guangzhou City, which are located in the Guangdong province. In addition to design services, LCC will also be providing high-level technical consulting services to assist Unicom in assessing its migration strategy and to help determine its technical objectives for other regions in the province.

Endwave Corp.—Has received a multimillion-dollar initial order from DMC Stratex Networks, a solutions provider for cellular applications and broadband wireless access. The order from DMC Stratex Networks calls for Endwave to design, manufacture, and deliver custom millimeter-wave transceivers.

FRESH STARTS

SignalSoft Corp. and SiRF Technology, Inc.—Announced a strategic partnership to provide an end-to-end location-services platform to wireless subscribers worldwide. With this agreement, mobile devices enabled by SiRF's SiRFStar GPS technology will interoperate with SignalSoft's Location Manager, providing wireless-network operators with the necessary location and gateway tools for delivering sophisticated location-based services to their subscribers.

Peregrine Semiconductor—Has opened a new design center in Melbourne, FL that specializes in high-reliability and radiation-effects product development.

Texas Instruments and ParkerVision—Announced a strategic partnership to produce advanced wireless chip sets for cellular phones, WLAN applications, and other next-generation wireless products.

Better On-Line Solutions (BOS)—Announced a worldwide sales and marketing alliance with Interactive Intelligence, Inc., a global developer of interaction-management software. Under the terms of the agreement, BOS will have access to the Interactive Intelligence reseller channel, and will be solely responsible for marketing, selling, and supporting the integrated solution.

Glassman High Voltage, Inc.—Launched their new, redesigned, reprogrammed website for customers and purchasers of high-voltage DC power supplies. The site, which is located at www.glassmanhv.com, includes a host of new features such as a product-locator pop-up menu from the home page that will enable searches by voltage or by wattage. Customers can review this information either in the HTML format or in a printed format, simply by downloading a PDF file. The redesigned website includes all of Glassman's 22 product lines and provides full specifications and mechanical drawings for each one.

RF Industries Ltd.—Announced that its RF Connectors Division has signed a national distribution agreement, covering its full line of coaxial connector and cable products, with Irvine, CA-based Surface Mount Distribution, Inc.

Transcat Calibration Services—Has opened a calibration laboratory in Foster City, CA—between San Jose and San Francisco—to serve the Northern California market.

Intercept Technology—Announced a joint marketing agreement for an interface between its PANTHEON PCB/hybrid/MCM design software and Quantic EMC's Omega PLUS signal-integrity and EMC simulator. The interface provides the ability to output a Quantic format NIF file, enabling users to invoke Omega PLUS and solve signal-integrity and EMC problems through analysis and simulation. The interface also allows PANTHEON users to benefit from the simulator's Terminations Editor tool to perform "what if" scenarios, which improve signal integrity, as well as reducing crosstalk and emissions.

Accelerated Technology, Inc. (ATI)—Has announced extended Nucleus RTOS support for the latest ColdFire® release from Motorola, the highly integrated MCF5272 microprocessor.

Decibel Products—Announced that its DB75Q4D08UT autotune combiner earned the UL Registration, an exclusive approval earned only by compliance with stringent safety standards. This combiner combines up to eight transmit signals in one output. MRF


GARNAOUI

Garnaoui Is Named As U.S. Cellular Engineering VP

HICHEM H. GARNAOUI has been named vice president of engineering at U.S. Cellular. Garnaoui joins U.S. Cellular from Nextel Communications, Inc., where he most recently served as vice president of network operations for the Mid-Atlantic region.

TDK Semiconductor Corp.—ANDY MILLS to senior vice president and general manager of the Broadband Communications Products group; formerly vice president of the Network Business Unit.

Directed Light, Inc.—JEFF SMITH to president and COO; formerly executive vice president and COO.

I-Bus/Phoenix—GUY THRAP to director of Power Products Development; formerly involved in directing hardware and firmware design, DFM, HIS, compatibility, and agency compliance at Powerware.

Crazy Eddie, Inc.—TED VAGELOS to president and COO; formerly president of Priceline Long Distance, LLC.

3DSP Corp.—HENRY MATHISON to vice president of sales; formerly vice president responsible for sales at Wyle Electronics.

The Cellular Telecommunications & Internet Association (CTIA)—JOHN WINDOLPH to senior vice president of business development; formerly executive vice president and chief marketing officer for the Lowe Lintas & Partners advertising agency. Also, KIMBERLY KUO to vice president of communications; formerly senior vice president for marketing and corporate communications at enfo Trust networks.

P-Com, Inc.—JAMES J. SOBCZAK to CEO; formerly president and COO.

Teledesic, LLC—MICHAEL MCGOWEN to general counsel; formerly a partner with the Perkins Coie law firm. Also, LEN QUADRACCI to vice president of system engineering; formerly director of system engineering.

Sitraka Mobility—SUJAN MENEZES to vice president of client services; formerly vice president of banking applications at 724 Solutions.

IPC—JOHN KANIA to director of government relations; formerly associate lobbyist with the Carmen Group.

TRILITHIC, Inc.—BRUCE G. MALCOLM to chairman and CEO; formerly president. Also, TERRY W. BUSH to president; formerly vice president of the Instruments Division.

Xemod, Inc.—SCOTT BEHAN to vice president of product development; formerly executive vice president of engineering at AML Communications.

APA Wireless Technologies—MITCHELL D. AURAN to president and COO; formerly director of strategic development at Cenetec.

Sirific Wireless Corp.—ROY GUNTER to CEO; formerly vice president and general manager of Siemens Wireless Infrastructure, Siemens Information and Communication Networks.

ITT Industries, Avionics Division—STEVE D'ONOFRIO to director of design engineering; formerly manager of digital design engineering.


D'ONOFRIO


BURNS

Narda Safety Test Solutions—KATHY BURNS to applications engineer; formerly associate engineer for RF safety products. MRF

SHORT COURSES

Network Analysis/Spectrum Analysis Measurements
October 16-19 (Wokingham, United Kingdom)
Agilent Technologies
Eskdale Rd., Wokingham, United Kingdom
Tracey Bull
+44 118-927-6741, FAX: +44 118-927-6862
e-mail: tracey_bull@agilent.com
Internet: www.agilent.com/find/education

Future Directions In IC and Package Design Workshop (FDIP)
October 27 (Royal Sonesta Hotel, Cambridge, MA)
Components, Packaging and Manufacturing Technology Society
Paul Baltes
(520) 621-3054, FAX: (520) 621-1443
e-mail: epd@engr.arizona.edu
Internet: www.cpmt.org/conf/fdip01/fdip.html.

Principles of Modern Radar
October 29-November 2 (Atlanta, GA)
Georgia Institute of Technology
Continuing Education
Atlanta, GA 30332-0385
(404) 385-3502
e-mail: conted@gatech.edu
Internet: www.conted.gatech.edu.

58th ARFTG Microwave Measurements Conference: RF Measurements for a Wireless World
November 29-30 (San Diego, CA)
Automatic RF Techniques Group (ARFTG)
IEEE Microwaves Theory and Techniques Society (MTT-S)
Dr. J. Stevenson Kenney, Technical Chair
School of Electrical and Computer Engineering, Georgia, Institute of Technology
Atlanta, GA 30332-0250
(404) 894-5170, FAX: (404) 894-4641
e-mail: jskenney@ece.gatech.edu
Internet: www.arftg.org

MEETINGS

Communications Design Conference
October 1-4 (San Jose Convention Center, San Jose, CA)
CMP Media, Inc.
(888) 239-5563
e-mail: info@commdesignconference.com

34th International Symposium on Microelectronics (IMAPS 2001)
October 9-11 (Baltimore Convention Center, Baltimore, MD)
IMAPS
Washington, DC 20002
(202) 548-4001, FAX: (202) 548-6115
e-mail: IMAPS@imaps.org
Internet: www.imaps.org

24th Annual Newport Conference On Fiberoptics Markets
October 15-17 Hyatt Regency Newport, Newport, RI)
KMI Corp.
Newport, RI 02840
(800) 343-4035, FAX: (401) 847-5866
e-mail: info@kmicorp.com
Internet: www.kmicorp.com

2001 IEEE GaAs IC Symposium
October 21-24 (Renaissance Harborplace Hotel, Baltimore, MD)
IEEE MTT-S, Electron Devices Society, and Solid State Circuits Society
2001 IEEE GaAs IC Symposium
Piscataway, NJ 08855
Mary Clemente
(732) 562-5350, FAX: (732) 981-1203
e-mail: m.e.clemente@ieee.org.

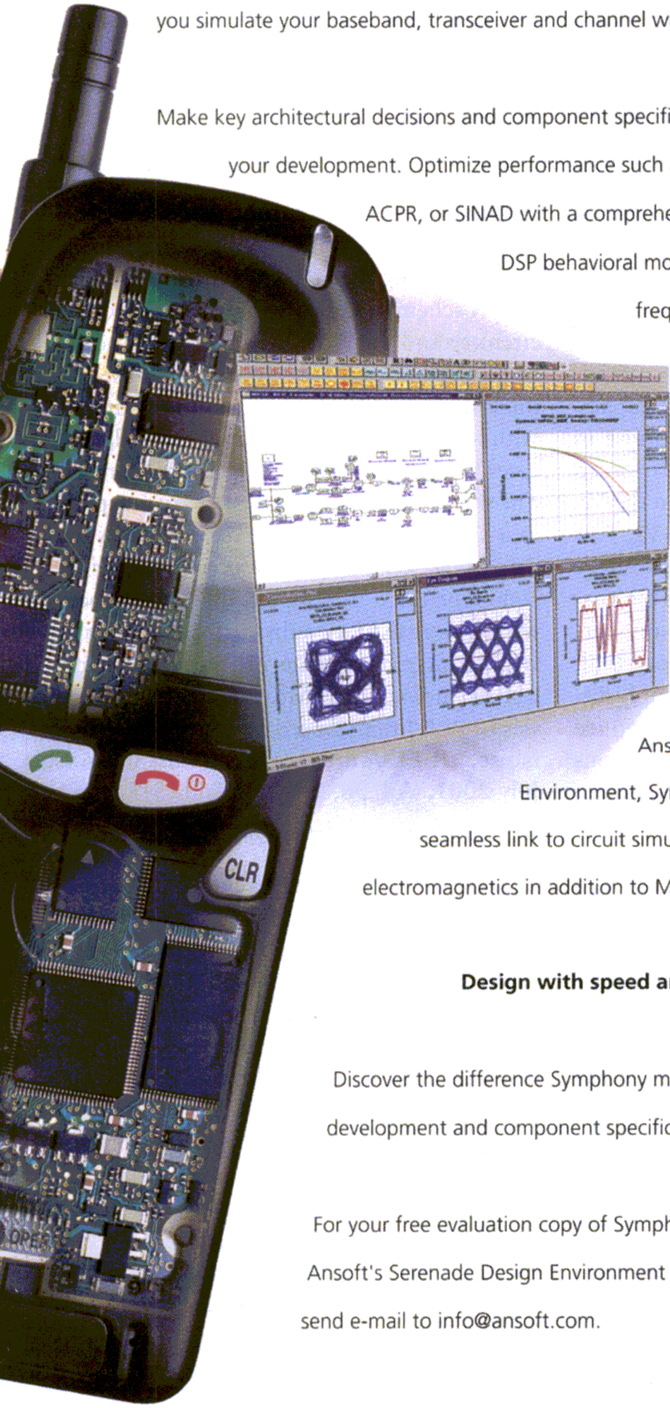
10th Topical Meeting on Electrical Performance of Electronic Packaging (EPEP 2001)
October 29-31 (Royal Sonesta Hotel, Cambridge, MA)
IEEE MTT-S Packaging and Manufacturing Society
George Katopis
IBM Corp.
(914) 845-6719, FAX: (914) 845-1593
e-mail: katopis@us.ibm.com
Internet: www.epep.com

3rd Advanced Technology Workshop on Packaging of MEMS and Related Micro Integrated Nano Systems
November 8-10 (San Jose Hilton South, Scotts Valley, CA)
International Microelectronics and Packaging Society (IMAPS)
Ajay P. Malshe
University of Arkansas/HiDEC
Fayetteville, AR 72701
e-mail: apm2@engr.uark.edu

CALL FOR PAPERS

IEEE MTT-S International Microwave Symposium
June 2-7, 2002 (Seattle, WA)
IEEE Microwave Theory and Techniques Society (MTT-S)
Technical Program Chair
Eric Strid, Cascade Microtech, Inc.
e-mail: eric@cmicro.com
Internet: www.ims2002.org
Deadline for technical paper summaries in .DOC format: November 26

14th International Conference on Microwaves, Radar and Wireless Communications (MIKON 2002)
May 20-22, 2002 (Gdansk, Poland)
Telecommunications Research Institute and Warsaw University of Technology Edward J. Sedek
Telecommunications Research Institute
+48 22 810 25 71
e-mail: godlewska@pit.edu.pl
Internet: www.pit.edu.pl/MIKON-2002
Manuscript deadline: December 15



CLR

Artificial neural networks model linear and nonlinear RF circuits

TYPICALLY, RF AND microwave circuit boards contain a large number of microstrip transmission-line discontinuities. To design the board properly, RF engineers must model these discontinuities accurately. Simulating them with conventional full-wave EM solvers can be very time-consuming, so simulators based on closed-form formulas are often used to accelerate the process. But these formulas involve the quasi-static approximation, are accurate only to a few gigahertz, and are restricted within certain parameters. One solution that has been used to overcome these limitations is to use EM models based on artificial neural networks, which can be coaxed to perform much faster and more accurately than EM solvers. Researchers S. Suntives, M.S. Hossain, J. Mittra, and V. Veremey of the Electromagnetic Communication Lab at Penn State University (University Park, PA) have devised a way to improve the effect of ANNs by using them to model the circuit's discontinuities rather than its S-parameters. They chose this approach because the frequency behavior of the lumped circuit parameters for a given range of physical and electrical parameters is much smoother than that of their S-parameters, and the equivalent circuit model is valid over a wider frequency range, which makes it easier to develop an ANN model of the circuit. See "Application of Artificial Neural Network Models to Linear and Nonlinear RF Circuit Modeling," *RF and Microwave Computer-Aided Engineering*, July 2001, Vol. 11, No. 4, pp. 231-247.

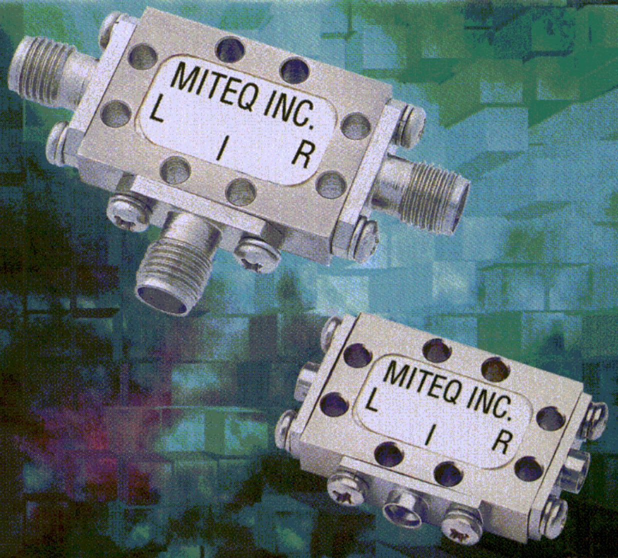
Study analyzes PLL jitter caused by digital switching noise

THE INTEGRATION OF analog and digital circuits onto mixed-mode chips reduces cost, but it introduces noise coupling that would not exist on either of the two types of circuits if they remained separate. For example, when an MOS transistor in the digital section of a mixed-mode chip turns on, it can discharge a capacitive load, generating a brief current pulse in the V_{ss} network. The current pulse is forced through the inductive bonding wire in the V_{ss} path and generates a voltage bounce on V_{ss}. This noise can couple through the resistive substrate or through a shared supply network to the analog section of the chip. Patrik Larsson, formerly with Bell Laboratories (Murray Hill, NJ), proposes using separate power-supply distribution networks in an attempt to reduce such noise coupling. His study investigates this type of noise coupling into a phased-lock loop, the most common analog block in today's mixed-mode designs. His work identifies the main noise-coupling mechanism for three different supply schemes. In the first scheme, the digital circuitry and the PLL share both V_{dd} and V_{ss}. In the second case, a separate V_{dd} is used for the PLL. In the third case, the PLL is supplied with separate V_{dd} and V_{ss} supplies. In all three cases, the local V_{ss} is used for substrate contacts. The three power-supply schemes are compared in a single chip containing several PLLs and digital noise generators. Measurements from circuits fabricated in a standard CMOS process with low-resistivity substrate are compared with results from circuits processed in a triple-well technology. See "Measurements and Analysis of PLL Jitter Caused by Digital Switching Noise," *IEEE Journal of Solid-State Circuits*, July 2001, Vol. 36, No. 7, pp. 1113-1119.

Researchers compare merits of different reduced-size resonant microstrip patches

RESONANT MICROSTRIP PATCHES can act either as antennas or as components of oscillators and filters in microwave ICs. In some applications (such as large arrays of antennas, cell-phone antennas, etc.), the patches are too large. An effective way to reduce the size of a patch resonating at a given frequency is to increase its electrical length by changing its shape. Several different shapes have been proposed. For antenna applications, rectangular ring patches, H-shaped patches, meander-shaped patches, and the insertion of multiple slots in a rectangular patch have been proposed. For applications requiring miniaturized coplanar-waveguide resonators, unilateral and bilateral comb-shaped patches have been proposed. Researchers G. Leon, R.R. Boix, and F. Medina of the University of Seville (Seville, Spain) have carried out a comparison of the relative merits of these designs. Using a numerical code based on the solution of an electric-field integral equation by means of Galerkin's method in the spectral domain, they present results for the resonant size and quality factors of the aforementioned patch designs fabricated on the same substrate at three different frequencies. See "A Comparison Among Reduced-Size Resonant Microstrip Patches," *Microwave and Optical Technology Letters*, May 5, 2001, Vol. 29, No. 3, pp. 143-146.

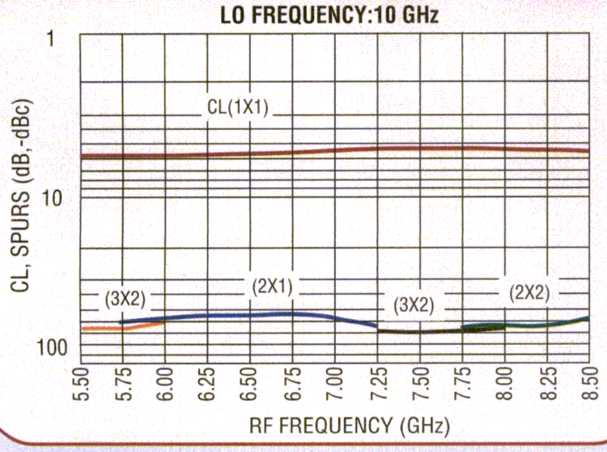
LOW SPURIOUS SPACEBORNE MIXERS



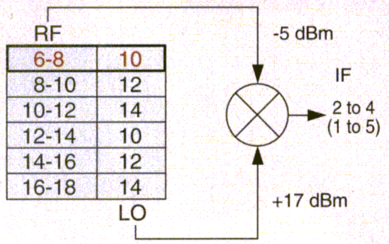
FEATURES:

- Broadband operation
- Minimal variation in conversion loss
- High IP3 and 1 dB compression versus LO power

CONVERSION LOSS/SPURIOUS



TYPICAL OPERATING BANDS



SPECIFICATIONS - Model TBR0618HA1/TBR0618HA1-S

RF/LO Input Frequency Range	6 to 18 GHz
IF Output Frequency Range	0.05 to 5 GHz
Conversion Loss	6 dB Typical
Spurious	-55 dBc
Third Order Intercept Point	+23 dBm Typical
1 dB Compression Point	+13 dBm Typical

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DESIGN

Simulation Tool Models And Verifies Timing Jitter In Oscillators

Simulation software allows designers to predict how random noise impacts phase noise and timing jitter at the circuit and system levels.

Phase noise and timing jitter are critical design considerations in nearly every type of digital communications system. In RF communications systems, low-noise oscillators are an integral part of phase-locked-loop (PLL)-based frequency synthesizers. Oscillator phase noise causes signal interference in nearby channels in the high-frequency spectrum, lowers the signal-to-noise ratio (SNR), and degrades the bit-error rate (BER) of a wireless transceiver. Oscillators are omnipresent in digital electronic systems that require clock signals for synchronization. In high-speed systems that employ PLLs for clock recovery, oscillator phase noise sets the limit on the overall jitter performance. As clock rates continue to escalate, minuscule variations in signal timing become even more crucial to system performance. Today, there is literally no room for error. That is why accurate phase-noise prediction and timing-jitter modeling are so important.

Phase noise and timing jitter are caused by the same underlying random phenomenon, and are nonlinearly related. Timing jitter is the characteristic associated with the random fluctuation in the transition times of a signal source in the time domain, while phase noise is the characteristic associated with the spectral purity of a periodic signal source in the frequency domain. Phase-noise spectrum is of principal interest

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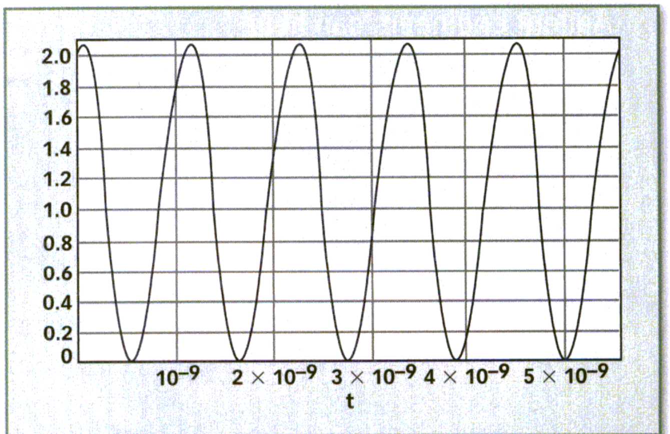
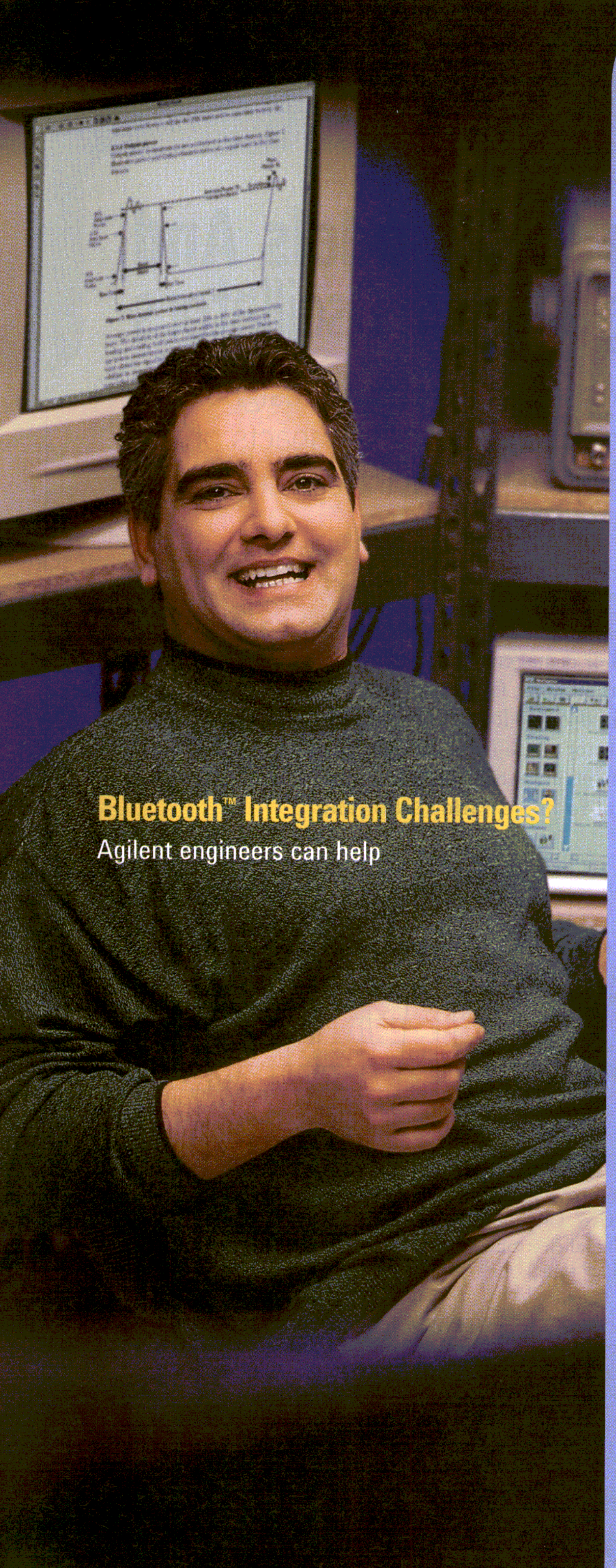


1. While no oscillator is perfect, this is what the waveform of a noise-free circuit would look like.



Bluetooth™ Integration Challenges?

Agilent engineers can help

It probably started with a conversation over in marketing: "*Bluetooth* wireless technology is the next big thing! We have to put it in all our products! Details? Bah, all you do is add an antenna. The engineers will figure it out. Let's go see if they're finished yet."

And now you and a few thousand other engineers are figuring out that *Bluetooth* integration is not a trivial task. From baseband DSP to RF interference, you've got an integration challenge worthy of legendary King Harald himself.

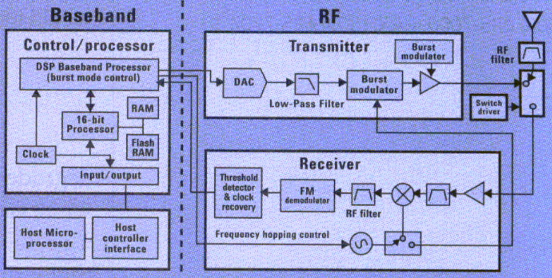
Welcome to the wild world of RF. New to RF? We've pooled the talents of our digital, DSP and RF experts to identify the most important signal checks you'll need to make when integrating *Bluetooth* designs. Our online resources include everything from an RF basics seminar to advanced measurement techniques.

Something for the RF experts, too. If you have the luxury of approaching *Bluetooth* from an RF background, we can offer advice on the most-efficient test procedures and toolsets to tackle a wide range of *Bluetooth* measurements.

The *Bluetooth* big picture. Most of the *Bluetooth* work we're seeing today involves the integration of a *Bluetooth* module into a new product design:

- Evaluating module performance and characterizing interoperability
- Understanding host-module integration issues
- Designing and debugging the host-module interface
- Conducting pre-qualification RF testing
- Getting *Bluetooth* Qualification
- Manufacturing quality products

Some of the more interesting problems show up in the second stage, as you bring the RF transceiver into your host products.



Watch out for some interesting interoperability problems when you integrate a Bluetooth module into your host device

Bluetooth™

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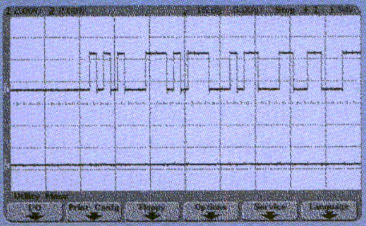
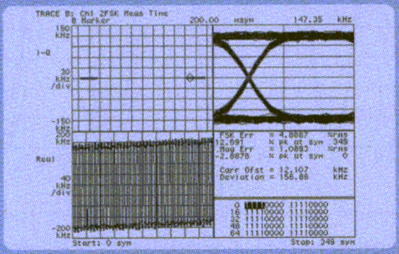
canada 1-877-894-4414, ext. 7471

Baseband signal integration. Challenges here include verifying transmission and receipt of data packets, viewing the actual data values transmitted, quantifying system bottlenecks, identifying logic errors, and resolving DSP and mixed-signal issues.

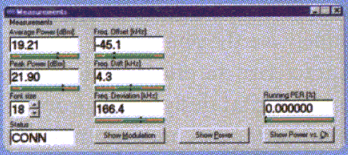
For instance, once you've found the preamble, you can identify the entire bit stream, including the access code, header and payload. Learn more in our free *Bluetooth* baseband application note.



The first two pulses in this idealized transmit signal correspond to the 0101 pattern of the preamble; the access code follows immediately after



Bluetooth measurement tools range from powerful design analysis to fast, automated tests for the production line. Above, a modulation characteristics test verifies proper performance of the modulation circuitry to ensure reliable data transfer over the Bluetooth communication link.



At left, an automated test combines pass/fail indications with numerical readouts

RF receiver tests. RF receiver performance is key to both *Bluetooth* qualification and overall product performance. For example, a sensitive radio that is immune to interference will reduce file transfer times and therefore increase battery life. You need to make sure the RF receiver will not be adversely impacted by the harmonics of high-frequency digital signals or other noise sources likely to be present in your system.

Receiver performance is tested in a number of ways for qualification, including carrier/interference and blocking tests. You probably won't need to run all the tests if you're integrating someone else's module, but they can be complicated so clear information and simplified procedures are important.

RF transmitter tests. The *Bluetooth* specification covers a wide range of transmitter tests, some to insure interoperability between *Bluetooth* devices (e.g., modulation characteristics) and others to meet regulatory limits (e.g., spurious emissions). Given the concerns about interference with other wireless systems, output spectrum tests are also important.

Integrating a module can create problems that affect transmitter performance, sometimes in unexpected ways. For example, power supply ripple coupled through your system can degrade the modulation characteristics.

You must be able to show that your device stays within both *Bluetooth* and regulatory limits, and the more of this work you can do on your design bench, the better. Some of the tests are complex and potentially time-consuming to understand and perform. Our free online application resources can help you look for and fix problems quickly.

Get the complete *Bluetooth* test story—FREE. Talk to one of our *Bluetooth* measurement specialists to learn more or tap into our free technical resources at www.agilent.com/find/bt.
Among the features you'll find there:

- A comprehensive measurement guide: *Performing Bluetooth RF Measurements Today* featuring descriptions and examples of the many RF measurements you might need
- Interactive measurements that show some key *Bluetooth* measurements in action, starting with frequency drift and frequency settling—explore these real-life measurements before you need to make them on your own system

We can't do much about your enthusiastic friends in marketing, but we can definitely help you through the *Bluetooth* integration experience.

Visit www.agilent.com/find/bt for a FREE *Bluetooth* CD-ROM packed with application notes, measurement tips and solution guides.

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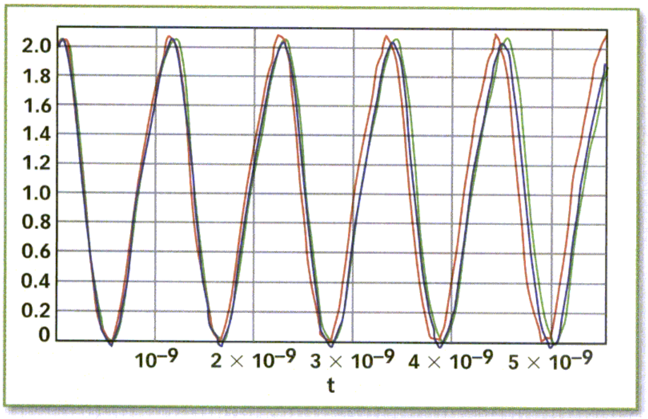
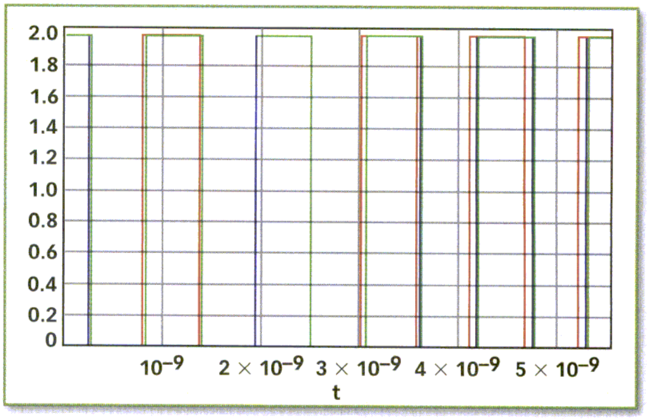


2. Unlike the oscillator in Fig. 1, a real-world oscillator exhibits phase and amplitude noise.



3. Timing jitter results in variations in the transition times of the waveform shown here.

in frequency-synthesis applications, whereas timing-jitter variation is a primary design specification in clock-recovery applications.

RF designers routinely use the phase-noise spectrum to characterize the effect of noise on oscillators. Digital designers prefer time-domain jitter to characterize precisely the same random phenomenon. This article outlines one workable solution for engineers designing PLLs, RF frequency synthesizers, and clock-recovery circuits. For these designers, oscillator timing jitter is an important hurdle to overcome. The article will demonstrate a method to bridge the gap between the RF world and the digital world, between phase noise in transistor-level circuit design and timing jitter in system-level design. This methodology will enable a designer to predict the impact of white noise and 1/f noise in devices on system-level phase noise and timing jitter.

The electronic-design-automation (EDA) tools used are Eldo, an RF integrated-circuit (RF IC) simulator and ADVance MS, a mixed-signal simulator, both from Mentor Graphics. Also part of the design methodology are hardware description languages (HDLs) which include Verilog, very-high-IC-HDL (VHDL), Verilog-AMS, VHDL-AMS, and SPICE/Eldo. VHDL and Verilog are the two most commonly used HDLs in the digital world. VHDL-AMS and Verilog-AMS are analog/mixed-signal extensions of these digital modeling languages. A language-independent simulator is very flexible and one that can simulate models written in all design languages.

Phase Noise and Jitter

In a Utopian world, a noiseless oscillator would provide a perfect time reference (a clock), because at steady state the time-varying oscillator waveform divides time into equal lengths. This simple concept **(Fig. 1)** shows the periodic waveform of a noise-free oscillator, a perfect time reference. In reality, however, all oscillators exhibit phase noise and jitter, nature's atomic clocks included. **Figure 2** shows what occurs when a free-running oscillator is perturbed by noise. This can be thermal noise, shot noise, 1/f noise in active and passive devices, or some other random disturbance. In each instance, noise causes amplitude and phase deviations.

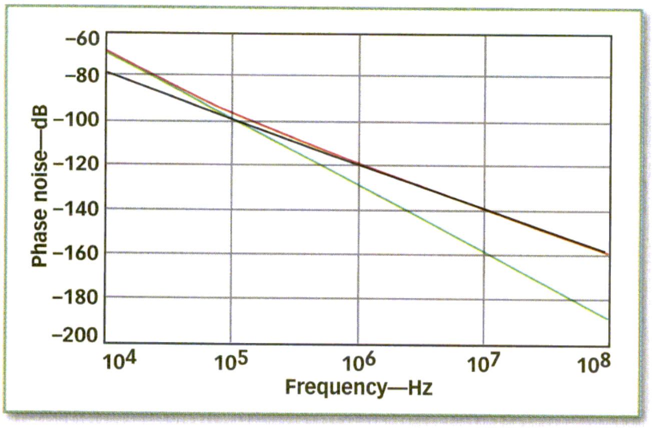
Phase deviation naturally accumulates with time and drifts without bound, because oscillators are autonomous circuits. After the phase of an autonomous circuit has been perturbed, the phase persists and cannot be restored without information from some other timing references. On the other hand, amplitude deviation always remains small and bounded, due to the fact that nonlinear oscillators are designed to operate around stable orbits.

Phase deviation and, to a lesser extent amplitude deviation, cause random variation in transition times and result in timing jitter **(Fig. 3)**. In PLLs, sequen-

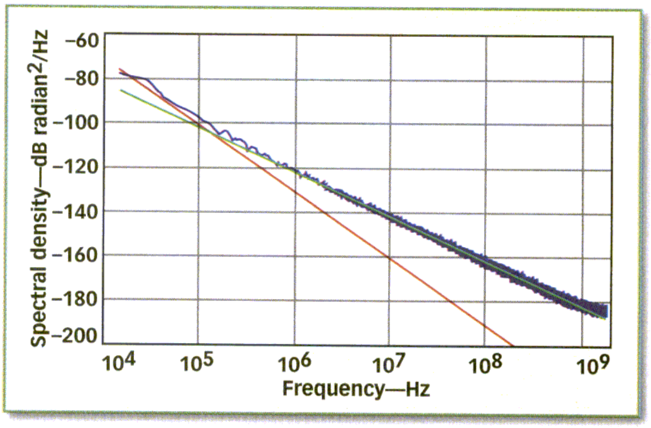
```
entity dosc_fd is
  generic (Freq   : real := 100.0e6;   -- oscillator frequency in Hz
           Ratio  : real := 1.0;       -- divider ratio
           TD     : time := 0 sec;     -- output delay
           Jdiv   : time := 0 sec;     -- sigma(jitter at one divider transition)
           Josc   : time := 0 sec;     -- sigma(cycle-to-cycle oscillator jitter)
           CFNosc : real    := 0.0;    -- oscillator phase noise due to 1/f noise
           Kcycle : integer := 10;     -- # of oscillator-divider cycle = 2 ^ Kcycle
           Seed1  : integer := 0;
           Seed2  : integer := 129792743);
  port (DOUT : out bit := '0');
begin
  assert Freq  > 0.0;  assert Ratio > 0.0;
end entity dosc_fd;

architecture bhv_jitter_fn of dosc_fd is
  constant halfPeriod : time := ( 0.5 * Ratio / Freq ) * sec;
begin
  process
    variable delta : time := 0 sec;
    variable state : bit  := '0';
    variable seed_1 : integer := Seed1;
    variable seed_2 : integer := Seed2;
    variable rn1, rn2 : real;
    variable rv3 : real_vector( 1 to 2 ** (Kcycle+1) );
  begin
    roscfn_vec( seed_1, seed_2, rv3, halfPeriod );
    for i in rv3'range loop
      rnorm( seed_1, seed_2, rn1 );
      rnorm( seed_1, seed_2, rn2 );
      wait for halfPeriod + sqrt(Ratio/2.0) * Josc * rn1 + (TD + Jdiv * rn2 - delta) +
           0.5 * MATH_1_OVER_PI * sqrt(CFNosc) * rv3(i) * sec;
      state := not state;
      DOUT <= state;
      delta := TD + Jdiv * rn2;
    end loop;
    assert FALSE report "roscfn_vec data exhausted" severity ERROR;
  end process;
end bhv_jitter_fn;
```

Fig. 4



5. This is the DSB phase-noise spectrum of the 1.82-GHz oscillator being simulated with an RF IC simulator.



6. The phase-noise spectrum plot shown here is generated from a time-domain jitter model (see Fig. 5).

tial phase detectors and digital frequency dividers sample phase noise and convert it into timing jitters. Note the cumulative nature of jitter with time. Phase noise and jitter are nonstationary random processes. The common assumption that phase deviation remains small is only valid over a limited time interval. This simplified analysis incorrectly predicts that the oscillator output spectrum contains ideal impulses, but this is inconsistent with the basic fact that all oscillators, natural or manmade, are imperfect time/frequency references. In truth, with the continuous presence of device noise, the phase deviation drifts without bound, and a single waveform is asynchronous with itself over a widely separated time interval. Over any time scale—short or long—phase noise is nonstationary. In a free-running oscillator, white noise causes timing jitter to increase with the square root of the time scale, while 1/f noise causes timing jitter to grow linearly with the time scale. Since device-level noise is always present and cannot be removed, characterization and modeling are key to keeping phase noise and timing jitter in check.

Time-Domain Jitter

A 1.82-GHz complementary-metal-oxide-semiconductor (CMOS) oscillator is used to illustrate the basic jitter principles. As a first step, the oscillator circuit was characterized using Eldo RF. From Eldo RF, the oscillator steady-state waveform and the phase-noise spectrum are obtained. From this, noise parameters can be extracted for a time-domain jitter model, which is simulated using ADVance MS, the language-independent mixed-signal simulator. Timing-accurate models of oscillators, along with the other PLL building blocks such as phase detector, charge pump, digital dividers, and analog filters are written using the Verilog, VHDL, Verilog-AMS, VHDL-AMS, and SPICE/Eldo HDLs. This allows an efficient analysis of an entire mixed-signal PLL and enables evaluation of system-level performances such as loop dynamics, timing jitter, and spurious sidebands.

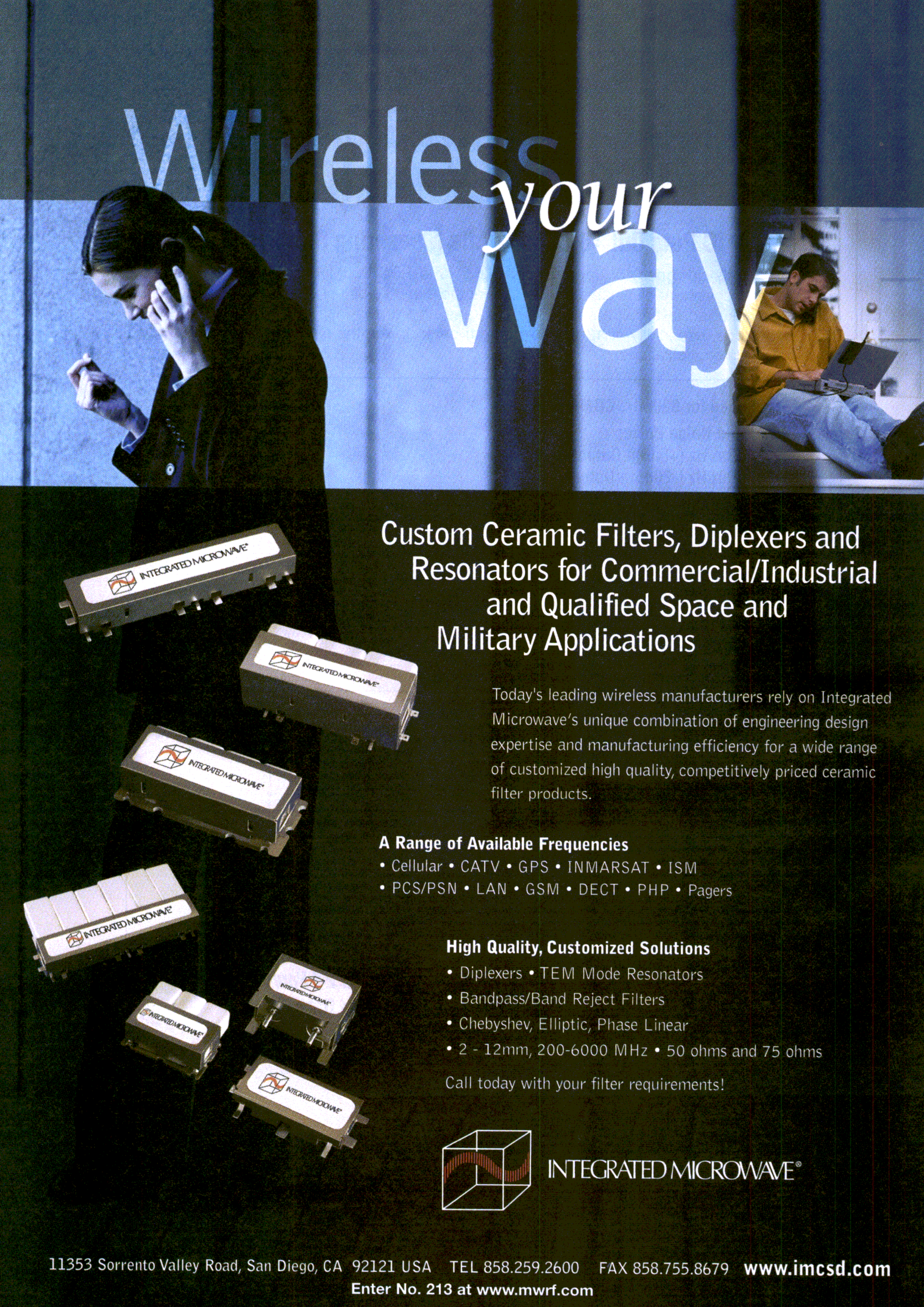
Taking a closer look, **Fig. 4** presents a VHDL model of an oscillator-divider circuit. Stationary, independent jitter that is due to divider amplitude-noise at the time of transition is modeled using a Gaussian random-number generator. Oscillator jitter that is caused by white noise (such as thermal and shot noise in devices) is modeled as a Gaussian random walk, which is the cumulative sum of independent Gaussian jitters. This is the discrete-time form of Brownian motion that was first observed by Robert Brown in the random, continuous movement of pollen grains suspended in water and was subsequently explained by Albert Einstein in 1905. Technically, Brownian motion is the continuous-time noise that is produced by time-integrating white noise that has equal amounts of random noise at all frequencies (colors). Oscillator jitter is the random walk obtained by a discrete-time sampling of Brownian motion. In the frequency domain, oscillator phase noise and timing jitter that is caused by white noise has a $1/f^2$ density.

Oscillator jitter that is due to 1/f device noise is more complicated. 1/f noise, which is also known as flicker noise, is random noise with magnitude that is inversely proportional to the frequency. Integrating 1/f noise over time produces $1/f^3$ phase noise. This noise is found very close to the carrier harmonics in an oscillator spectrum. Time-domain $1/f^3$ noise modeling is complex because this type of noise is nonstationary, self-similar, and correlated with long-term time dependence.

In the time domain, this random noise, like Brownian motion, looks similar when magnified at different time scales and has long-range correlation. In the frequency domain, the phase-noise spectrum is the same over many decades of frequency. Contrary to common belief, oscillator phase noise cannot be modeled as a stationary noise source, such as a sum of sinusoids with random phases. In the VHDL model of Fig. 4, DSP techniques are used to integrate 1/f noise while preserving the inherently nonstationary and self-similar nature of $1/f^3$ phase noise.

The accuracy of the phase-noise spectrum obviously determines the accuracy of the time-domain jitter model. The 1.82-GHz CMOS oscillator was simulated using the RF IC simulator, and produced a (one-sided) double-sideband (DSB) phase-noise spectrum 2L(fm) **[Fig. 5]**. The predict-

Continued on page 90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Design Of Short-Range Radio Systems

The first installment of this four-part series offers design techniques and regulatory issues in advanced short-range radio systems.

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Short-range radios, also known as microradios, have existed for several decades, primarily in the form of one-way control and security-class links. With an increase in integration and processor control, the design of these radios has become more of a system than a circuit issue. The design of these microradios is further complicated by regulatory issues, such as which carrier frequency to use, choice of modulation scheme, whether to use transmit-power averaging, and the type of antenna for a particular design. In addition, cost issues include receiver (Rx) topology, frequency-source topology, synchronization format, level of integration, baseband processing, and when to step up to two-way links or to move to industrial-scientific-medical (ISM) frequency bands. Part 1 of this four-part series will review the basics of radio-wave propagation, while Parts 2 and 3 will cover regulatory- as well as system-oriented issues and design methodologies, respectively.

Microradios are commonly associated with consumer applications such as remote-keyless-entry (RKE) devices for automobiles and garage-door opening systems. Bluetooth represents the high end of the product range identified by the generic term "microradio." These "control-class" applications have historically been one-way systems, sometimes so cost constrained as to feature on-off-keyed (OOK) inductive-capacitive (LC) or surface-acoustic-wave (SAW)-based transmitters (Txs) that consist of little more than a single transistor oscillator that is modulated by keying its power supply with an encoder chip that can perform key-press detection and some form of rudimentary encoding. Low-cost Rxs have included an LC or SAW regenerative Rx, a topology that can be implemented with only a few transistors. Digital control has been added in recent years, to a level often featuring baseline microcontrollers such as the Microchip PIC12C509A or the Microchip KEELOQ [code-hopping encoders from Microchip Technology, Inc. (Chandler, AZ)]. Power supplies for portable-radio units now typically consist of one or two lithium (Li) coin-cell batteries.

With the availability of higher-frequency, cost-effective complementary-metal-oxide-semiconductor (CMOS) and bipolar-CMOS (BiCMOS) semiconductor processes, microradio technology is moving toward higher levels of integration. With more powerful digital control, these radio systems are poised to move beyond control applications and into network data communications and wireless data acquisition

(DAQ). As the complexity of these short-distance radio systems increases, engineers must apply standard wireless-system design techniques such as the use of a link budget.

A link budget considers Tx power, path loss, antenna gain, and Rx sensitivity when calculating radio range. The link budget is not meant as an exact calculation, but to provide desired reliabilities as a function of range and operating conditions. The difference between radio range under ideal free-space conditions and in an environment with more realistic signal degradation can be more than an order of magnitude. A suggested approach is the use of a second-order model that uses a path fade which is higher than inverse square, and the assumption of a log-normal probability distribution of signal strength with standard deviations ranging from 4 to 16 dB as a function of environment.

The mathematics for this level of link budget is simple, and will be presented here in a way that is also applicable to certification testing, where analyses are made of field strengths some distance from the device under test (DUT). A derivation can start with the effective aperture of the receive antenna, which is the ratio of the power delivered to the load to the incident RF power density. Effective aperture can be thought of as the area where a 100-percent efficient antenna captures all of the energy that would otherwise pass through the same area without the antenna. The maximum effective aperture is related to directivity, D_0, the maximum directive gain of an antenna on its main lobe and wavelength, λ, by:

$$A_{em} = \lambda^2 D_0 / 4\pi \quad (1)$$

The directivity does not take into account losses due to mismatch and ohmic losses, so the effective aperture, A_e, is equal to eA_{em}, where e is the total efficiency. For a perfectly isotropic (omnidirectional) antenna without losses, $D_0 = 1$. The closest practical antennas to this performance are quarter-wave whips and similar designs. A quarter-wave whip shows a directivity of approximately 1.7 and efficiency losses exclusive of matching of generally less than 1 dB. The gain of the antenna varies as a function of relative orientation which, for mobile terminals, is not well-controlled and must be viewed statistically. An acceptable practice for a particular microradio application is to measure path loss at various antenna orientations and positions relative to the human body that are appropriate for that application, come up with an average loss relative to an isotropic antenna, and then lump antenna-gain variation as a function of posi-

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tion into the standard deviation of path loss.

Effective aperture can be used to convert the root-mean-square (RMS) field strength at the antenna into power delivered to the Rx input:

$$P_{rec} = (E_{rms}{}^2 / \eta)A_e \qquad (2)$$

where:

η = the impedance of free space (377 Ω).

E_{rms} = the RMS field strength at the antenna, and

P_{rec} = the power delivered to the Rx input.

Since the power levels permitted by the Federal Communications Commission (FCC) are presented in terms of field strength, this relationship is handy for measuring fundamental and harmonic signal levels. European regulations are based on units of effective radiated power (ERP), or the power that would be radiated from a perfectly isotropic antenna which matches that received on the peak of the actual antenna's main lobe.

Note from Eq. 1 that A_e is dropping for a particular antenna type such as quarter-wave whip as the inverse square of frequency. From Eq. 2, it can be seen that if electric field is constant over frequency with A_{em} dropping over frequency, then P_{rec} must be declining with the inverse square of frequency. This is usually referred to as increasing path loss with frequency, a somewhat confusing choice of terminology, since this loss occurs even if power density is frequency independent. What is actually physically happening is that the ability to gather the power density is declining over frequency if directivity (receive-antenna type) is held constant. It is as if a smaller lens is being used to focus sunlight. This fact must be accounted for in regulatory harmonic measurements—the "free" 6-dB/octave drop due to the increase in free space path loss versus frequency (with scaled antennas) must be taken back out to calculate the field strength of harmonics correctly. The only way to hold constant or increase A_e with increasing frequency is to introduce a larger and directional antenna.

Receive power for a particular transmit power over a free space link is provided by the Friis Transmission Equation. For polarization-matched antennas that are aligned on directionality maximums this equation reduces to:

$$P_r / P_t = \left\{\left[(\lambda / 4\pi)^2\right]\right. \quad R^n G_{0t} G_{0r} \qquad (3)$$

where:

P_r = receive power,

P_t = transmit power,

R = range (in meters),

n = the path-loss exponent (2 in free space),

G_{0t} = the gain of the transmit antenna, and

G_{0r} = the gain of the receive antenna.

These gains are the same as directivity multiplied by efficiency loss. For practical link calculations, it is helpful to massage Eq. 3 into a form giving range as a function of degrading factor "D" (the linear form of all decibel losses in a practical link from ideal), Rx sensitivity S (milliwatts are most convenient), and transmit power Pt (the same power units as S). These manipulations yield:

$$R_{max} = \left[(c / 4\pi f)^2 (DP_t / S)\right]^{1/n} \quad (4)$$

When converting from ERP to field strength, as is done in comparing US and European regulations, several other relations come in handy. The power density, S_r (in watts per square meter) of a uniform plane wave is provided in terms of RMS electric-field strength; free-space impedance, η; and effective radiated power, P_{terp}, as:

$$S_r = E^2 / \eta = E^2 / 120\pi = P_{terp} / 4\pi R^2 \quad (5)$$

The last term follows from radiated power and the area of a sphere of radius R. From this equation, it is possible to find RMS field strength, E_{RMS}, at range R in meters (ideal inverse-square propagation) and transmitted isotropic effective radiated power, P_{terp}, as:

$$P_{terp} = 0.03333 R^2 E_{rms}^{\ 2} \quad (6)$$

$$E_{rms} = (5.477 / R) (P_{terp})^{0.5} \quad (7)$$

This basic compliance-oriented physics flows directly into link budgeting by taking degrading factors into account as shown in Eq. 4. An excellent source of raw data specifically for the 900 MHz ISM band is ref. 1. These data may be expected to remain approximately true for losses in the 300-to-500-MHz range, normally used for control and security applications. Depending on environment (such as indoor or outdoor, building type, range, operation between floors, etc.), the path-loss exponent changes from 2.0 for free space to a range from 1.8 to 5.0.

Safety Margin

It is also true that the received signal strength may be approximately modeled for reliability purposes as log normal, meaning that it shows a Gaussian distribution (in decibels) over a large number of samples. The standard deviation of this signal-strength variation will typically vary from 4 to 16 dB over a wide range of operating conditions. A few days of engineering time invested

in properly modeling the link statistics for a particular application will pay enormous benefits in optimum system design. To use this information in product specification and system design requires the addition of a safety margin to the link budget to provide the desired reliability. This safety margin is most conveniently specified as a number of standard deviations in the statistical variation of path loss (in decibels), with a deliberately selected reliability at the maximum range. A brief example along the lines of a garage-door opener may be illustrative.

For example, assume that a Tx is operating at 416 MHz under FCC 15.231 rules (which will be reviewed in Part 2) with a transmit ERP of −15 dBm. Television harmonic interference is assumed negligible. The selected Rx shows a noise figure of 8 dB, a bandwidth of 60 kHz, and a demodulation and forward-error-correction (FEC) combination that requires 12 dB of final signal-to-noise ratio (SNR) to achieve the desired bit-error rate (BER). The Rx sensitivity is calculated to be −106 dBm. The mean transmit-power degradation due to antenna orientation and body absorption is experimentally determined to be −10 dB. Experimentation also shows that under the desired operating conditions, the link displays a path-loss exponent of 2.5 and a standard deviation in signal strength of 7 dB.

Desired Reliability

It is desirable to determine effective maximum range for a 95-percent chance of a successful transmission. From any table of a normalized Gaussian distribution, it can be seen that 1.65 standard deviations will have an area of 0.9505 under the density curve. In order to achieve the desired reliability, 1.65×7.00 dB = 11.60 dB is added to the link losses, providing a total required link-loss safety margin of 21.6 dB, or a degrading factor D = 0.00692. Plugging these numbers into Eq. 4 yields a 95-percent reliable range of 61 m. Reviewing the graph on p. 108 of ref. 2 shows a 99-percent reliability for any random range from 0 to 61 m (the service area). The range of this same link under free-space conditions can be calculated at approximately 2000 m, a range which would never hold up in practice.

Next month, this three-part article series will continue with an examination of regulations for short-range radio systems in the US and in Europe. MRF

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Use A Sampling Power Meter To Determine The Characteristics Of RF And Microwave Devices

This two-part article discusses triggering, DAQ, and peak-mode measurements using a sampling power meter.

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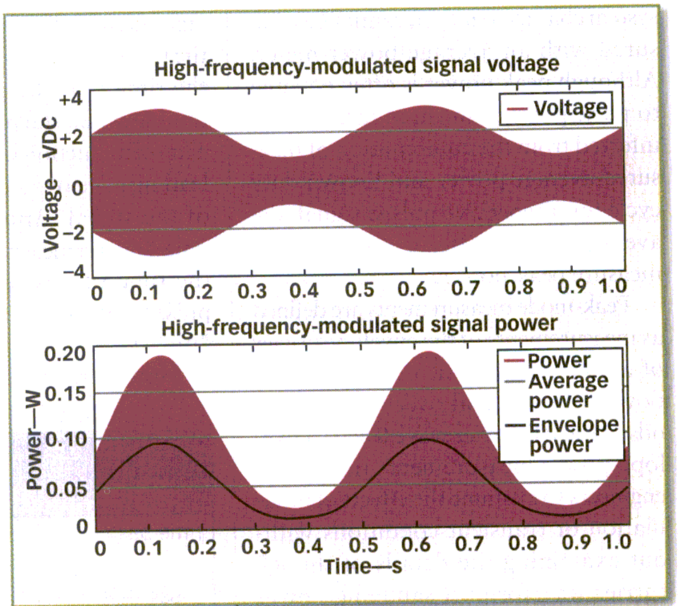
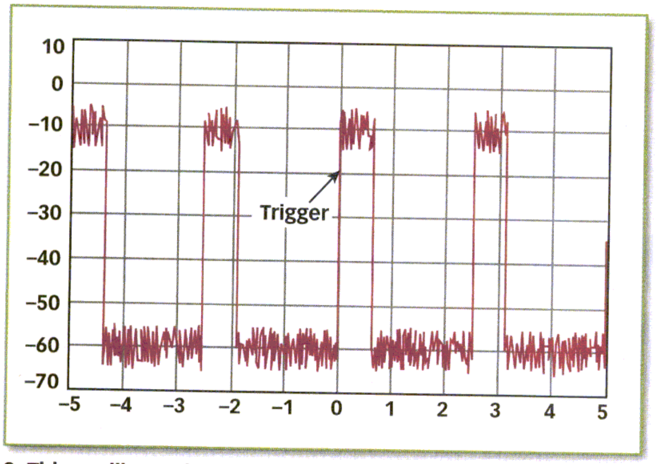
Modern communications systems are placing increasingly greater demands on RF design and manufacturing engineers. Finished systems need to be tested to ensure compliance to exacting international standards. Component parts are increasingly tested using the same types of complex signals as their target application. Wireless-system designers now require full characterization of RF components and subsystems under dynamic signal conditions before they can qualify these devices for use in their designs. And the old stalwart of RF measurements, the standard power meter, is creaking under the strain of these new requirements.

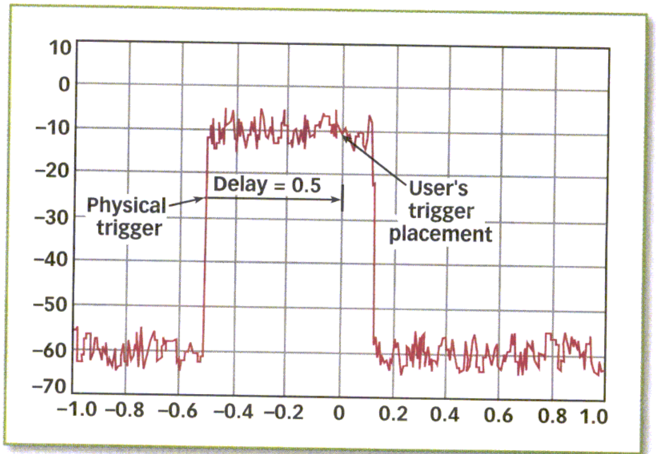
This two-part article outlines the advantages of using a sampling power meter to determine the characteristics of RF and microwave devices, using the digital sampling oscilloscope (DSO) for comparison. Part I discusses the DSO, presenting its capabilities from the user's perspective and reviewing developments in DSO technology. Part II will cover the sampling power meter, presenting its capabilities from the user's perspective and reviewing technological developments. Sample measurements will com-



1. These illustrations show the power envelope of a high-frequency-modulated signal.



2. This oscillograph shows a typical triggered display.



3. This oscillograph shows a trace display with trigger delay.

plete the article. The most accurate way to quantify an RF signal is to use an average-power meter. RF power will continue to be the standard measure by which any RF signal is quantified, verified, and referred back to national transfer standards. Unfortunately, the measurement requirements of modern communications systems go beyond the capabilities of this instrument. For example, peak power, which is an increasingly important measurement for safety and system characterization, cannot be measured with an average-power meter. Although peak power is often equated to pulse power, which, in turn, can be inferred from the combination of measured average power and known duty cycle for a repetitive pulsed signal, the average-power meter cannot directly measure peak power.

Peak-mode measurements are defined as measurements of the envelope power of an RF signal, which is the average power of the signal over several periods of the RF carrier wave **(Fig. 1)**. Envelope-power measurements thus allow engineers to examine the effects of modulation or transient conditions without examining the details of the RF carrier waveform. A sampling power meter is capable of performing peak-mode measurements. It can provide RF engineers with answers to questions such as: What is the maximum power output from this device? What is the pulse droop? What is the maximum output power before the signal's peak-to-average ratio is reduced? Does the output signal meet the Global System for Mobile Communications (GSM) power-mask specification?

To answer these questions, the sampling power meter must have some features that are not present on the average-power meter. The first essential feature is triggering. Triggering allows the user to make measurements in the time domain, relative to some known event (i.e., the trigger). For example, the average power in an RF pulse can be calculated most accurately by computing the mean of the envelope power during the time that the pulse is on. Determining when the pulse is on is usually a known function of time relative to the leading edge or some other time stamp of the signal. Another example of a time-domain measurement is rise time—the time that is required for the RF pulse to transition from an "off" to an "on" state.

However, not all signal types have a regular time stamp that permits time-relative measurements. Nor will a user always require this measurement. But one can still derive important measurements from sampling and acquiring the envelope power. Peak power, average power, and peak-to-average ratio all can be calculated without the need for a trigger.

The second distinguishing feature of a sampling power meter is the speed at which it can track, sample, and measure the RF signal's envelope power.

The simplest method of adding peak measurements to a power meter is to change the detector architecture to sense the peak power rather than the average power. As this architecture only enables the measurement of peak power, it does not answer all of the questions previously outlined. For that reason, this discussion concentrates on instruments that are capable of various peak-mode measurements. These instruments exclu-

Table 1: DSO display controls

CONTROL	UNITS	DESCRIPTION
Time base	Seconds per division	Controls the time span of the displayed waveform
Trigger position	Left, center, right	Controls where the trigger instant is displayed
Trigger delay	Seconds	Allows the user to effectively select the trigger event position at some arbitrary point in time. This is useful when the physical trigger event (perhaps an external signal) is synchronous, but not coincident, with the points of interest in the displayed waveform.
Amplitude range	Volts per division	Sets the y-axis scaling
Input offset	Volts	Offsets the center of the display

Table 2: Oscilloscope settings

Time base	1 per division
Trigger position—left, center, right	Center
Trigger delay	0
Amplitude range	10 per division
Input offset	30

sively use diode sensors, followed by sampling-acquisition architecture.

During their development, sampling power meters have taken two paths. One path was to combine the power-detection diode output with a digital oscilloscope. The second path was to speed up an average-power meter to such an extent that it delivers peak-mode measurements. But upon examining the functional requirements of a sampling power meter, it is clear that neither of these architectures are optimal. To offer some common ground with engineers who are unfamiliar with sampling power meters, the following discussion refers to the capabilities of the DSO. It also provides a basis for comparing and contrasting the two devices.

Triggering allows the user to make measurements that are time-relative to the trigger instant. This section describes the controls that users need to take advantage of this characteristic. **Table 1** shows the traditional display controls that a DSO offers.

These controls may not always appear under display controls, but they do control the appearance of the displayed waveform. **Figure 2** shows a scope-type display, and **Table 2** shows its settings.

With these controls, the user can inspect the measured waveform anywhere within the limits of the instrument. For example, the signal shown in Fig. 2 can be viewed as shown in **Fig. 3**

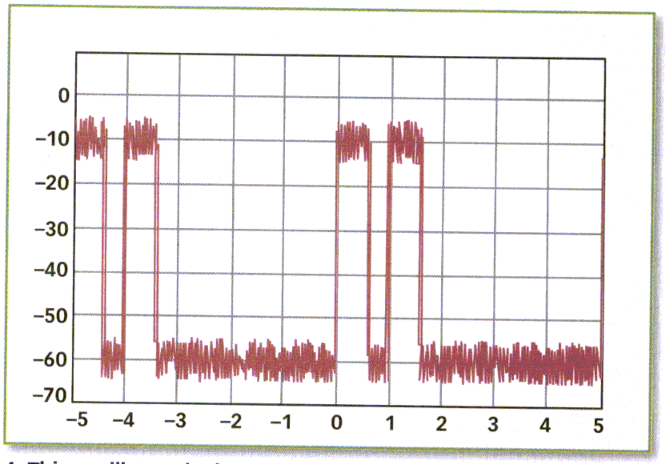
Table 3: Alternative settings

Time base	0.2 per division
Trigger position—left, center, right	Center
Trigger delay	0.5
Amplitude range	10 per division
Input offset	30

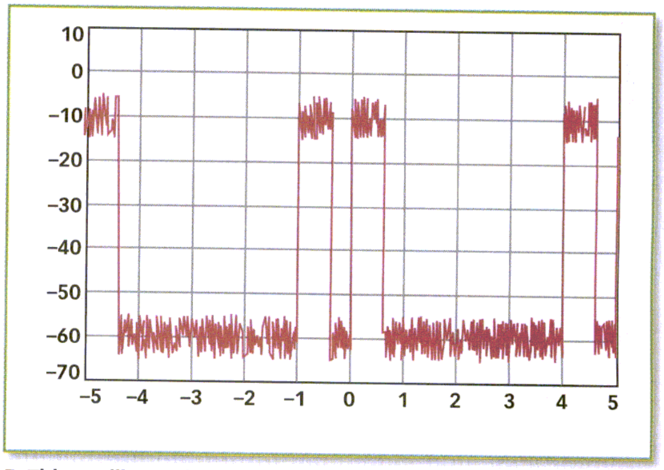
and as described in **Table 3.**

Thus, the user can control where the signal appears within the display by varying the controls described before **(Table 4).**

The purpose of this article is not to explain the operation of the trigger circuits of digital oscilloscopes. However, it is worth introducing some of the concepts conventionally applied within these instruments. They commonly have a settable comparator circuit that changes state whenever the input signal crosses a particular level. The instrument stores the time at which that event occurred to allow all of the input-signal samples to be tied to the trigger instant. The trigger circuit generally runs at all times. The acquisition system digitizes and stores the input signal at the achievable rate. If the acquisition system had stored the required number of samples to display the waveform at the current display settings and a trigger occurs, then the acquisition is stopped and the sampled data are transferred to the display. The acquisition system then has a little "dead" time to enable data transfer before restarting.



4. This oscillograph shows a double-pulse signal.



5. This oscillograph shows a double-pulse signal triggered on the second rising edge.

This strategy works well in many situations. Consider the waveform displayed as an oscilloscope trace **(Fig. 4),** assuming the same settings as for Fig. 2.

Since the acquisition system and trigger run asynchronously, it is possible to trigger on either of the rising edges. So, the oscilloscope display may appear as shown in **Fig. 4** or **5**, or, more likely, as shown in **Fig. 6**. This situation can be avoided by using the trigger hold-off function.

Trigger hold-off stops the trigger system from providing an output every time the input voltage crosses the trigger level. The action of the trigger system is "held-off" for a user-settable time interval after every trigger event. In the double-pulse example, if the hold-off time is set to 2, the trigger system will be disabled for the second of the two pulses. The trigger is then returned to the stable triggered display shown in **Fig 4.**

This is a form of trigger filtering or qualification. Various filters—highpass, lowpass, or bandstop—are often switched into the signal path before the trigger to help provide stable triggers. Advanced digital oscilloscopes can also provide more sophisticated trigger qualification, such as triggering, if the pulse is greater than or less

Table 4: Trigger controls

CONTROL	SETTINGS/ UNITS	DESCRIPTION
Trigger source	External Channel 1 Channel 2	This enables selection of the trigger source. Thus, the instrument can be triggered from one of the measurement channels or from an external source. The external trigger input is usually not displayed and normally has a restricted functionality
Trigger mode	Normal Auto level	This determines whether the trigger level is set up manually by the user or automatically by the instrument
Internal level	Volts	This allows the trigger level to be set anywhere on the current display
Delay	±seconds	This delays the displayed trigger from the actual trigger. Allows the user to effectively select the trigger-event position at some arbitrary point in time. This is useful when the physical trigger event (perhaps an external signal) is synchronous, but not coincident, with the points of interest in the displayed waveform
Trigger hold-off	Seconds	This allows the trigger system to be stopped (held off) for a length of time
Slope	Rising or falling	This determines whether the trigger system looks for a low-to-high or high-to-low transition

than a set time interval.

Continuing with the DSO example, it is commonplace for these types of instruments to offer manual and automatic measurements that go beyond counting the graticules on the display.

Table 5 defines some of the measurements featured on most DSOs.

Technology Overview

The major challenges in designing a digital oscilloscope are to get the signal from the probe to the display as fast as possible, and to make the display behave similar to an analog oscilloscope. Anything beyond that could be considered "nice-to-have" features. The oscilloscope is also a very good time-measurement instrument, so anything that helps preserve the time accuracy is a worthwhile investment.

The acquisition system for a digital oscilloscope can be described generally as a set of switchable gain amplifiers followed by a fast analog-to-digital converter (ADC). The ADC resolution usually is not more than 8 b, but its sample rate is extremely fast. The voltage range that the ADC must cope with

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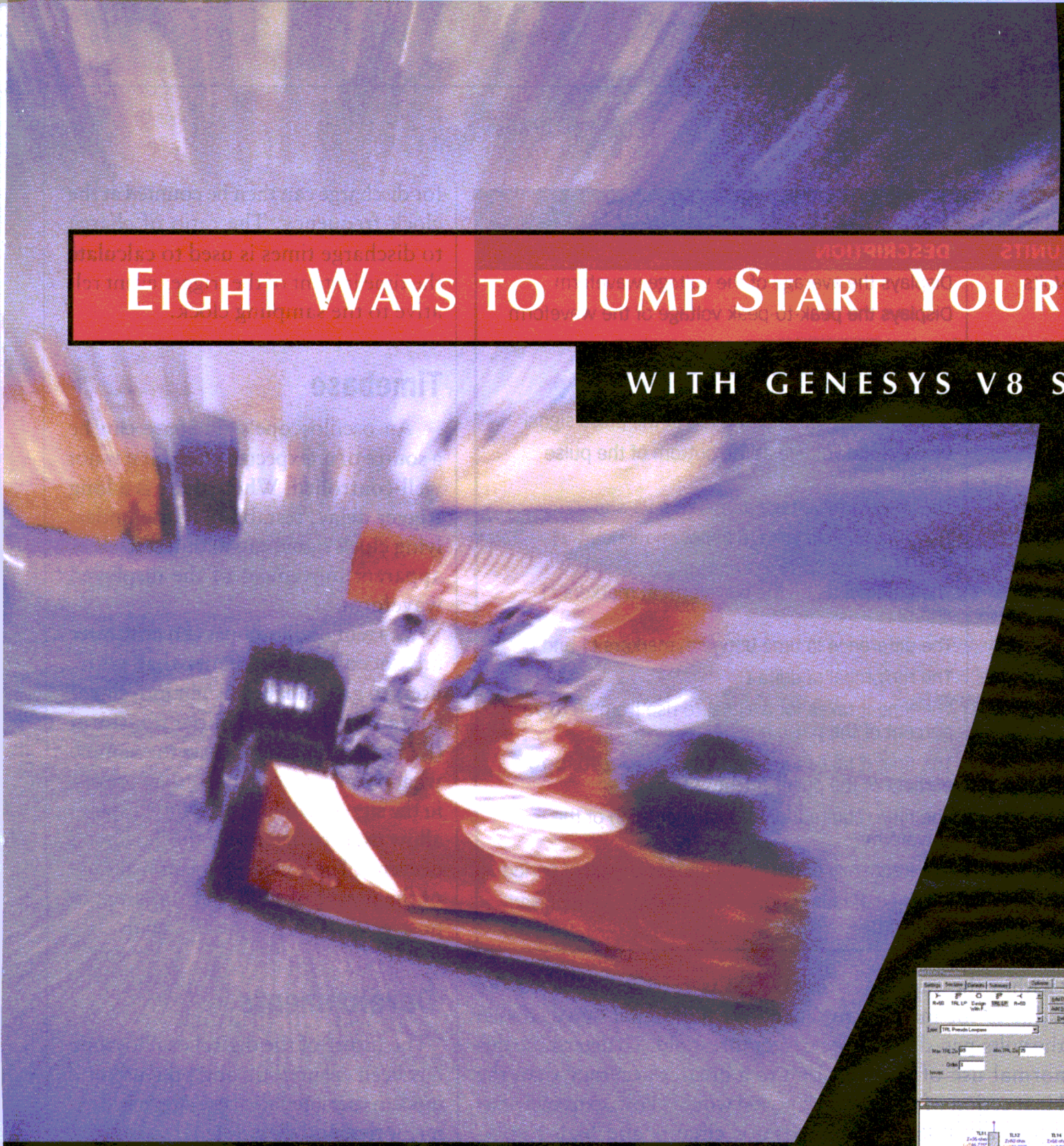

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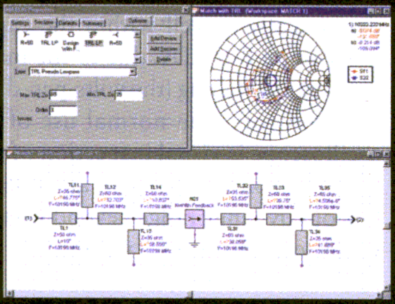

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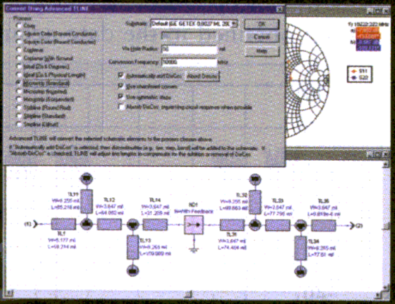

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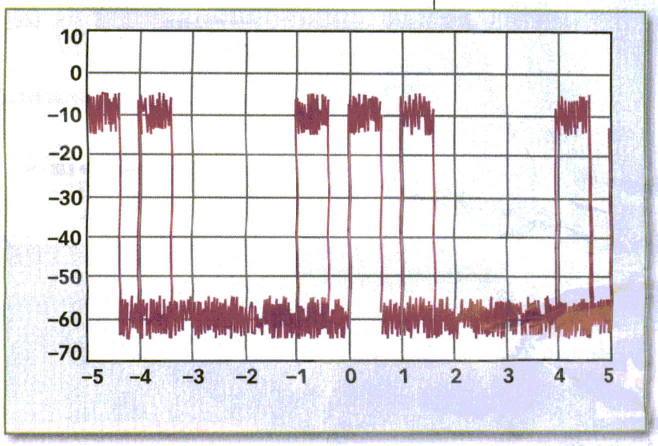
Table 5: DSO measurement definitions

MEASUREMENT		UNITS	DESCRIPTION
Voltage	Average	Volts	Displays the average of the voltage waveform
	Peak-to-peak		Displays the peak-to-peak voltage of the waveform
	RMS		Displays the RMS calculation of the voltage waveform
	Pulse top		Displays the voltage at the top of the pulse signal
	Pulse bottom		Displays the voltage at the bottom of the pulse signal
Voltage marker	V1, V2	Volts	The waveform voltage at markers V1 and V2
	Delta V		The difference in voltage between markers V1 and V2
Time marker	t1, t2	Seconds	The time points on the waveform at markers t1 and t2
	Delta t		The difference in time between markers t1 and t2
	1/delta t	Hertz	The reciprocal of delta t
Time	Rise time	Seconds	The time it takes for a signal to rise from 10 to 90 percent of the pulse height
	Fall time	Seconds	The time it takes for a signal to fall from 90 to 10 percent of the pulse height
	Period	Seconds	The time required for one full repetition of the waveform
	Frequency	Hertz	The frequency of the signal (the reciprocal of period)
	Duty cycle	Percentage	The ratio of high-to-low pulse intervals for a pulsed waveform

dynamically is well within the capability of an 8-b ADC due to the manual range switching (the volts-per- division control). The normal use of an oscilloscope is visual inspection of the waveform on the screen, so any instantaneous dynamic range beyond what is visible on the screen is wasted.

Trigger

This is an extremely important part of the oscilloscope. All of the timing information obtained in the data acquisition (DAQ) is referred back to the instant at which the input signal crossed the trigger threshold. To determine this instant to a greater accuracy than the clock period would allow, extremely fast comparator circuits and high-resolution pulse-stretching techniques are used. With a common implementation of this type of circuit, the time between the (asynchronous) trigger instant and the sampling clock is captured as charge on a capacitor. The capacitor is then discharged with a longer time constant than its charge-time constant. The time for discharge can then be counted at the clock frequency. The ratio of charge to discharge times is used to calculate the time instant of the trigger event relative to the sampling clock.



6. This oscillograph shows a double-pulse signal with indeterminate triggering.

Timebase

An oscilloscope's timebase can be a source of unexpected effects if it is not well-controlled. Where there is a long trigger delay, instability in the underlying clock signal can manifest itself as apparent movement of the displayed waveform.

Current oscilloscopes can now offer advanced measurements that allow users to gain greater insight into signal characteristics. Fast Fourier transform (FFT) analysis allows users to examine what frequency components are present in the signal, while probability displays allow users to determine the likelihood of events. These measurements make use of the power of modern processors to post-process the acquired signals.

Developments

The future of the digital oscilloscope has been ensured through the use of a modern user interface, fast sampling, deep memories, and digital signal processing (DSP). The latest oscilloscopes have intuitive user interfaces with online help, tutorials, and the new perspective of probability. Modern electronics have the ability to process signals so fast that waveform display rates have reached a point where it is difficult for the user to obtain further information from the instrument. Probability-enhanced displays can provide the user with more information by highlighting the parts of a repetitive waveform with the highest probability, while showing the low-probability excursions. Thus, the user can quickly obtain a picture of how a circuit is performing.

The second part of this article will expand the discussion to cover the underlying technology to help the reader understand a sampling power meter. Sample measurements will complete the article. MRF

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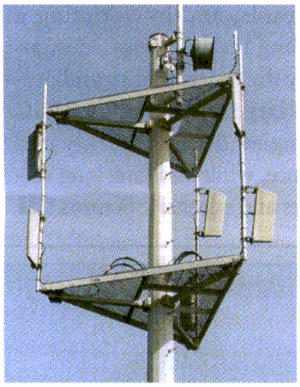

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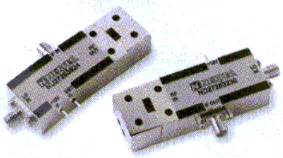

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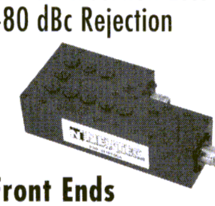

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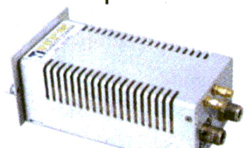

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Continued from page 70

ed spectrum was in good agreement with silicon (Si) measurements. $1/f^3$ phase noise is dominant at sideband frequencies that are below 100 kHz, while $1/f^2$ phase noise is dominant above this frequency. Integrating white noise over time produces 1/f (squared) phase noise, Brownian motion. This noise is found around the carrier harmonics in an oscillator spectrum.

The self-similar nature of $1/f^3$ and $1/f^2$ noises are obvious from examining the phase-noise spectrum. The oscillator was then simulated by using a Mentor Graphics Corp. (Wilsonville, OR) time-domain jitter model in the language-independent simulator. Spectral analysis of the oscillator transition times resulted in the phase-noise spectrum that is shown in **Fig. 6.** This is in agreement with the expected spectral density. By using HDL modeling, the gap was closed between phase noise in the transistor-level circuit design and system-level timing jitter.

The Simulation Challenge

PLLs and oscillators are complex, nonlinear dynamic systems. In the past, designers ran long transistor-level PLL simulations that took weeks. Not so long ago, designers relied on their intuition and crafted specialized C programs to analyze PLLs more efficiently. Even then, the effect of white noise and 1/f noise on oscillator phase noise and timing jitter was rarely modeled. The relationship among device-level noise, phase noise, and timing jitter is not obvious, particularly where 1/f device noise and 1/f3 phase noise are important. Nonetheless, some progress has been made recently, although some challenges lie ahead.

Today's revolution in high-bandwidth communications technologies has created a demand for mixed-signal and system-on-a-chip (SoC) solutions. The growing design complexities make it imperative for engineers to shift their design methodology to meet the challenges ahead. This article describes a Bottom-Up methodology proceeding from transistor-level circuit design to system-level verification. There is no reason why this process cannot be reversed, going from system-level design to transistor-level implementation. In a Top-Down methodology, new ideas are first tested and debugged at the architectural level, and system-level budgets, such as jitter, are determined using timing-accurate models and mixed-signal simulation. With the new generation of simulation tools, and by adopting a HDL modeling methodology, designers can verify the functionality and performance of large and complex RF IC and mixed-signal designs quickly and with confidence, while avoiding long verification cycles and expensive Si turns. MRF


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Comparing Integer-N And Fractional-N Synthesizers

This review of two popular frequency-synthesis techniques helps system designers choose the best approach for their communications applications.

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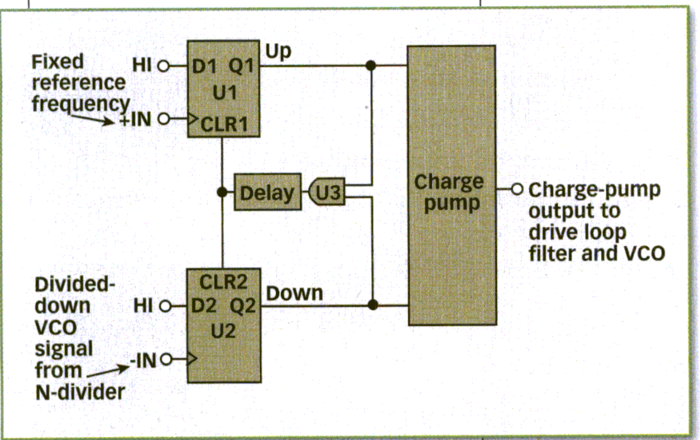
Fractional-N synthesizers for phased-locked-loop (PLL) applications have threatened to challenge the dominance of integer-N synthesizers for many years. These synthesizers offer the dual advantages of significant improvement in PLL phase noise and fast lock times. Due to this, PLL designers betray their excitement with each new fractional-N product. Still, the integer-N frequency synthesizer is more wide-ly used for local-oscillator (LO) signal generation. It may be possible to explain the popularity of the integer-N synthesizer by reviewing the theory and fundamentals of each technique, and examining the basic building blocks for each approach. The advantages of each technique will be reviewed to provide system designers with some direction in choosing which method is better for a particular application. The cellular GSM-900 system will serve as the example to provide numerical examples of how integer-N and fractional-N synthesizers work.

The first step is to examine the workings of an integer-N PLL, which has a number of basic building blocks, including a phase-frequency detector (PFD), charge pump (CP), a loop filter, a voltage-controlled oscillator (VCO), and a feedback divider. The PFD is the engine of any PLL, comparing the phase (and frequency) of two inputs, while its output drives the charge pump to control the loop filter and VCO. The two inputs are a fixed reference frequency (IN+) and the divided-down, fedback signal of the VCO (IN−). This is the essence of any negative feedback system. (All negative feedback-control theory transfer functions are applicable to PLLs.) If the IN− phase lags IN+, the PFD drives the VCO to increase its frequency. This will continue until the IN− signal phase is the same as the IN+ phase. This is how phase lock is attained. If the IN− signal phase leads the IN+, the opposite occurs as the PFD drives the VCO to decrease in frequency again—until lock is achieved. Since phase is the integral of frequency, a similar frequency-domain



1. The PFD and charge-pump circuitry in an integer-N synthesizer tune the loop filter and VCO according to detected phase and frequency differences at the two input ports.

analysis can be used to explain how frequency lock is attained **(Fig. 1)**.

The PFD works in exactly the same way for integer-Ns and fractional-Ns. When in lock mode, the PFD still produces very narrow alternating high and low-output pulses that maintain lock and prevent deadbands. The PFD is also the dominant inband noise source in the PLL. Commercially available synthesizers differ in phase-noise performance mainly due to the noise characteristics of their PFD. For any particular synthesizer, as the PFD frequency increases, the PLL will become noisier. The reason for this is that if the PFD is updating the VCO at a faster rate, it is also going to contribute more noise. The noise performance degrades at a rate of 10log (PFD frequency).

The loop filter, which is usually a passive element, is situated between the charge-pump output and the VCO. It is employed in a lowpass fashion to attenuate noise and spurious elements of the LO. Since loop-filter theory does not vary when applied to both synthesis techniques, this part of the PLL is not a differentiator between the two approaches.

The VCO generates an output frequency that is dependent on input voltage. Most VCOs have positive polarity, meaning that an increase on the tuning-port voltage corresponds to an increase in its output frequency. Its output is split, with part of the signal providing the LO signal to the outside world, and the remaining portion providing the feedback signal into the N-divider of the synthesizer.

The N-divider takes the output signal from the VCO and divides it by a preprogrammed amount. This divided signal is fed to the IN– input of the PFD. A combination of dual-modulus bipolar prescalers and complementary-metal-oxide-semiconductor (CMOS) counters is the most widely used methods of implementing N-dividers. The N-divider is the ratio of the RF frequency to the PFD frequency:

$$N = RF_{OUT} / FPFD$$

As its name suggests, an integer-N synthesizer only allows integer values of N to be used in this equation. This limits the freedom of the RF output. In GSM-900, the RF output must be able to move in channel steps of 200 kHz. This means that using an integer-N forces PFD frequency (FPFD) to also be equal to 200 kHz. Therefore, to attain an RF output frequency (RF_{OUT}) of 900 MHz, N is programmed to divide by 4500. To tune to the next adjacent channel, located at 900.2 MHz, N is programmed to 4501.

Since it is based on integer-N, the fractional-N PLL inherits many of the building blocks of its predecessor **(Fig. 2)**. The PFD, charge pump, loop filter, and VCO all work in the same way on both platforms. The N-divider is different, however. In a fractional-N PLL, the N-divider is broken up into the integer divider (N) and a modulus-M interpolator (M), which acts as the fraction function by toggling the N-divider. The interpolator is programmed with some value (f). The average division factor is now N + f/M where:

$$0 < f < M.$$

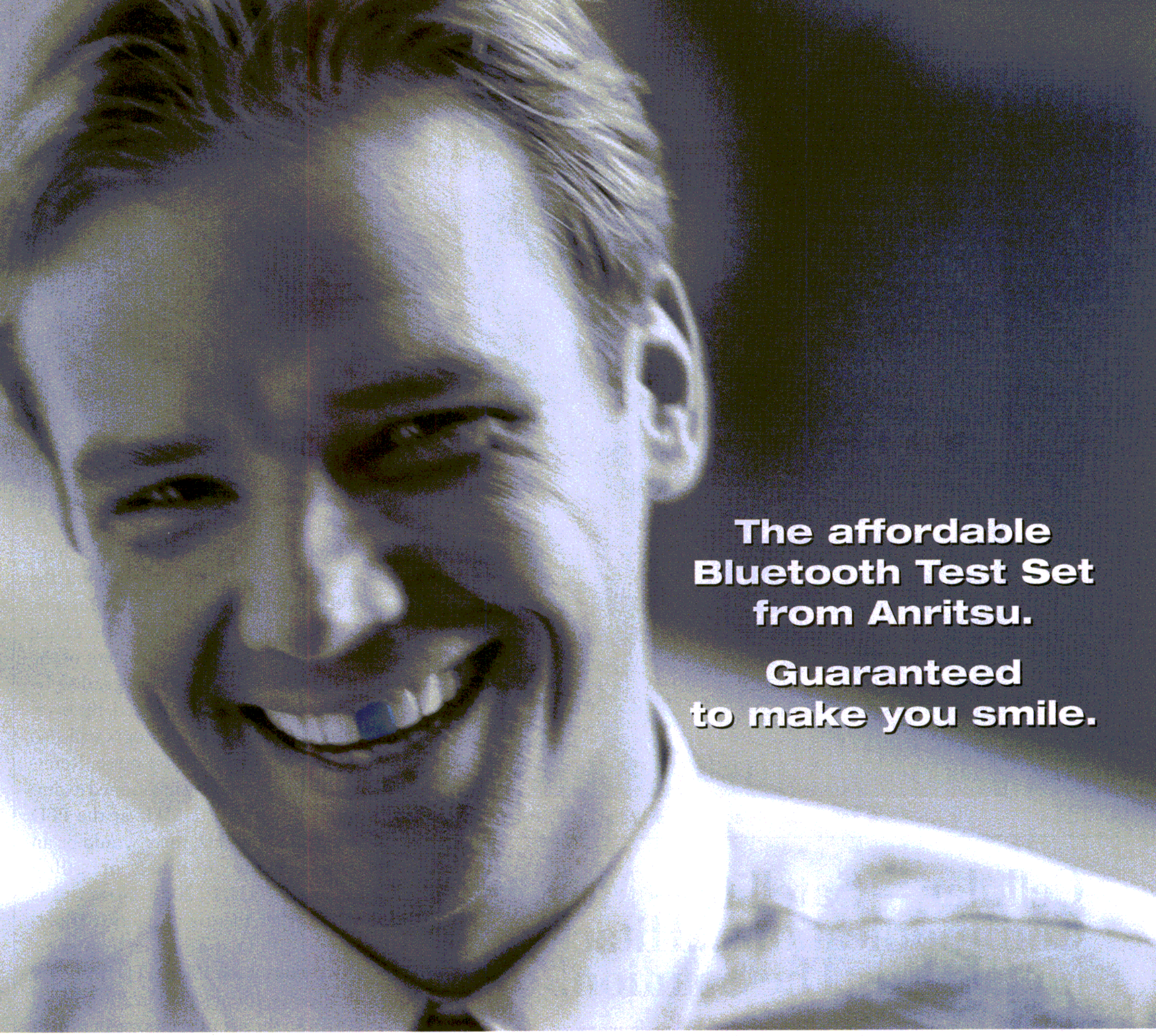
$$(N + f/M) = RF_{OUT} / FPFD$$

This is the essence of fractional-N synthesis. It now means that the PFD frequency can be larger than the RF channel resolution. In revisiting the GSM-900 example, it may be instructive to examine how the fractional-N approach handles the generation of 900-MHz output signals with 200-kHz channel resolution. If a modulus M of 10 is available, F_{PFD} can be set to 2 MHz. N is programmed to 450, f is 0, and M is 10. To tune to 900.2 MHz RF_{OUT}, $N_{AVERAGE}$ must be 450.1, N is programmed to 450, f is 1, and M is 10. To achieve this, the N-divider is toggled under the control of the interpolator between N and N+1 and the average taken. What effectively occurs is that the N-divider divides by 450 nine times, and then divides by 451 once every 10 PFD cycles. The average over the 10 cycles of 450.1 is taken as $N_{AVERAGE}$, which is fed to the PFD. However, much complex circuitry is needed to implement this.

Interpolators can be implemented using the overflow bit of an accumulator. Alternatively, sigma-delta modulators are often employed for this task due to their averaging function and noise-shaping characteristics. In this case, every time an N value is presented to the PFD, it has been modulated by the sigma-delta modulator. This introduces spurs to the loop at F_{PFD}/M. This modulation of the N-divider introduces inaccuracy into the loop in the form of phase error. In the case of integer-N, the loop inherently attains lock, and the PFD minimizes phase error. In a fractional-N, the sigma-delta modulator ensures the average phase error is correct, since it regularly introduces phase error into the loop. This makes the PFD and charge pump work harder to compensate and maintain lock. This brings the linearity of the charge pump into the equation. All charge pumps have some nonlinearity associated with them. Nonlinearity in the charge pump will increase the spurious amplitude.

Using a second-order sigma-delta modulator, the spurious noise on the charge-pump output appears in a first-order highpass nature **(Fig. 3a)**. However, the spurious signals are then acted upon by the lowpass loop filter before being presented to the VCO **(Fig. 3b)**. This is why second-order sigma-delta modulation is the method that many PLL designers choose. The spurious content is introduced in a highpass fashion, and then attenuated by the lowpass filter.

Since using higher-order sigma-delta modulation will noise-shape at a higher order, it is at the expense of current consumption, more spurs, and greater design complexity. Other standard sigma-delta tricks, such as dithering, can be used in fractional-N. This reduces the spurs at the expense of noise degra-



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dation **(Fig. 4)**. Spurious content is the major drawback to fractional-N, and the effect of spurious noise on a system will be discussed later.

As shown for integer-N synthesis, the N-divider can be quite large. The phase-noise performance of a PLL is a measurement of PFD noise floor that is degraded by 20logN. Since N is 4500 in the GSM-900 setup, 20log4500 is 74 dB. Therefore, since N is so large, the RF output is degraded by 74 dB.

While the integer-N solution has worked extremely well since its invention in the 1930s, a question arises as to when superior phase-noise specifications are required, as inevitably happens. From a phase-noise standpoint, the need for fractional-N is an obvious one. As mentioned previously, the phase noise of a PLL is a measurement of the PFD noise floor that is degraded by 20logN. If N could be reduced, the overall degradation due to 20logN will also reduce, improving phase noise. Referring to the previously mentioned example, if there were 2 MHz at the PFD instead of 200 kHz, there would be an improvement in noise of 20log (2MHz/200 kHz) = 20dB. However, as noted earlier, there is a penalty for operating the PFD at a higher frequency. This penalty is the 10log (PFD frequency increase) = 10log (2 MHz/200 kHz) = 10 dB. So the net phase-noise improvement is 20 – 10 = 10 dB. This 10 dB represents a very significant reduction in noise, but there are drawbacks—having 2 MHz at the PFD means that the RF output can only increment in steps of 2 MHz. For Global System for Mobile Communications (GSM), the requirement is to increment the RF output in steps of 200 kHz. Using a fractional-N with a modulus M of 10 would reduce the noise, while maintaining the RF output resolution of 200 kHz. The sigma-delta modulator can introduce noise into the loop. In practice, the noise introduced into the PLL by the sigma-delta modulator can sometimes be greater than the improvement gained by using a lower value of N. Good design practice can prevent this from happening. The ADF4252 fractional-N syntheiszer from

ADI shows typical inband phase-noise performance of −100 dBc/Hz when set up in GSM-900 conditions. This is typically 8 dBc/ Hz better than the ADF4113 integer-N.

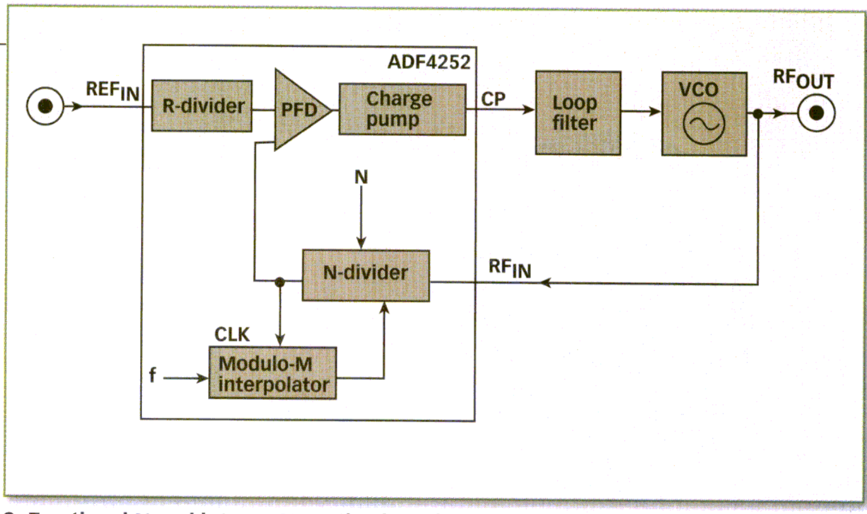
Lock time is another reason for pushing the cause of fractional-N synthesis, especially in mobile communications. Faster lock time will improve a handset's battery life. If it takes the handset less time to reach power-up conditions, search for a signal, and return to power-down conditions, it will consume less power. GSM base stations often employ two integer-N PLLs that are set up in a "ping-pong" architecture since they cannot lock quickly enough by themselves. These architectures prove extremely expensive, requiring an extra VCO and high-isolation switches to prevent interference between the two PLLs. A fast-locking PLL would eliminate the need for two PLLs, thus drastically reducing cost and size. There are two rules of thumb for PLL design that need to be considered:

- Faster lock times are achieved with increasing PFD frequencies. The reason for this is that the PFD updates the control voltage to the VCO at a frequency equal to the PFD frequency.
- The loop-filter closed-loop bandwidth (CLBW) is usually designed to be one-tenth (at most one-fifth) the PFD frequency. This is common practice to ensure loop stability and to realize realistic filter components.

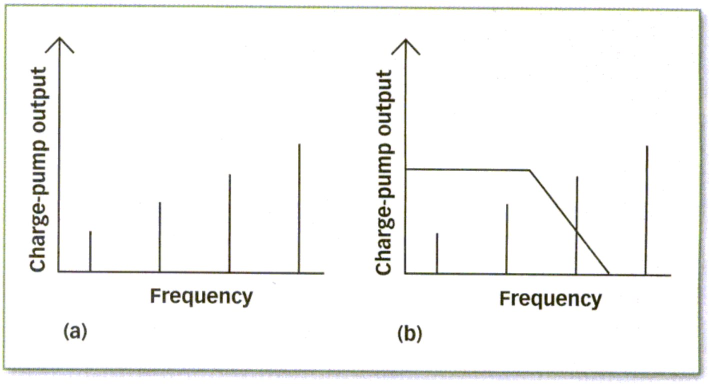
For GSM, an integer-N synthesizer is forced to have 200-kHz PFD and, therefore, the CLBW of the filter will typically be 20 kHz. This results in a lock time of approximately 250 μs. Increasing the CLBW is possible, but will lead to stability issues. A PFD of 2 MHz and CLBW of 200 kHz would lock more quickly—approximately 20 μs. It would seem that having a larger PFD is a good way to reduce lock time, and to improve phase noise. So, why then is the integer-N PLL not dead yet? The answer lies in how the fractional-N is implemented.

Any spurious noise on the RF output is not welcome, since it can lead to interference with other channel users. Naturally, stringent specifications exist on this. Spurious content in a PLL is generated mainly by the charge pump. Even when in lock, the charge pump constantly updates the voltage being sent to the VCO. This is to compensate for charge-pump leakage current and to prevent backlash. As the PFD obtains an input at IN+, it compares it to IN−, and sends an output to the VCO. Therefore, the DC voltage being fed to the VCO will have a pulse modulated onto it with a frequency equal to IN+. This modulation finds its way onto the output of the VCO, and manifests itself as a spur on the RF output at a frequency equal to IN+ on either side of the carrier. Harmonics of the spur will also appear on the RF output. The loop filter will generally have a cutoff frequency of approximately one-tenth the PFD frequency. In the case of an integer-N synthesizer, this means that a 200-kHz spur will have one decade of attenuation

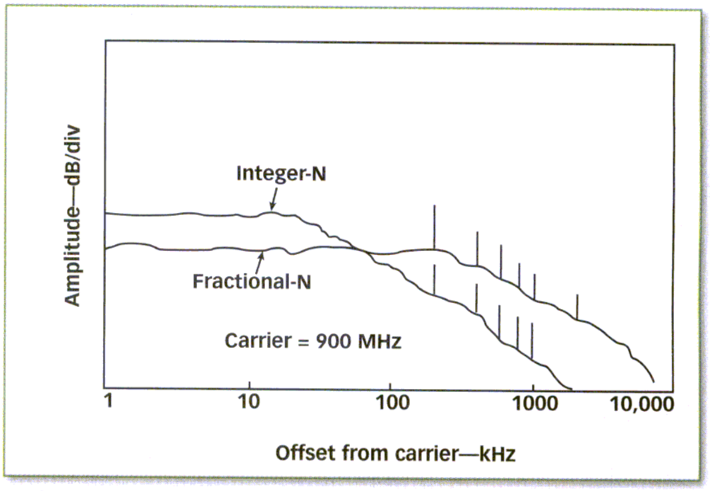
Continued on page 133



2. Fractional-N and integer-N synthesizers have many common elements, but the N-divider operation is different in each type.



3. With a second-order sigma-delta modulator, the spurious noise on the charge-pump output appears to have a first-order highpass nature (a), which is then acted upon by the lowpass loop filter (b).



4. This plot compares spurious levels for the two frequency-synthesis approaches.

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Selecting Prescalers For PLL Synthesizers

High-frequency prescalers are available in packaged and chip formats to provide the necessary frequency-division function for RF and microwave frequency synthesizers.

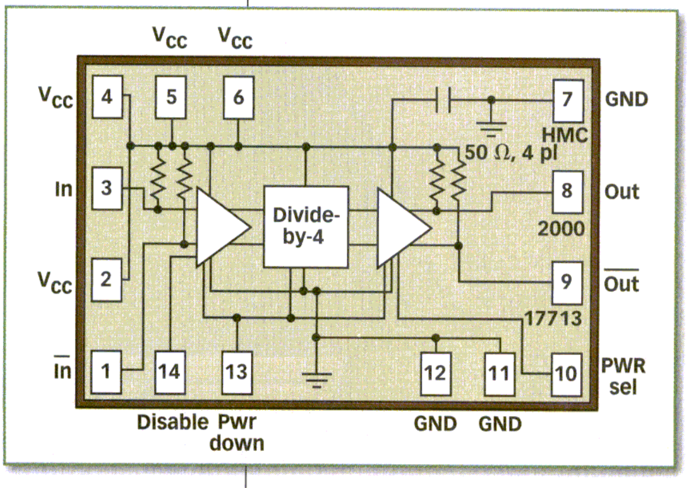
Prescalers are frequency dividers that are used in RF and microwave frequency translation and signal generation. They are commonly employed in phase-locked loops (PLLs) and frequency synthesizers to match the frequency of a high-frequency source to that of a reference oscillator. By understanding how to bias and operate prescalers, they can be effectively designed into a wide range of high-frequency applications.

DESIGN STAFF
Hittite Microwave Corp., 12 Elizabeth Dr., Chelmsford, MA 01824; (978) 250-3343, Internet: www.hittite.com.

The modulus or divide ratio is a fundamental prescaler parameter. Fixed modulus prescalers that have a modulus equal to an integer power of 2 (i.e., 2, 4, 8...) enable the highest input-frequency-handling capability compared to other integer division ratios. For example, a new line of monolithic-microwave-integrated-circuit (MMIC) prescalers from Hittite Microwave Corp. (Chelmsford, MA) features high input-frequency capabilities to 13 GHz. The prescalers, which are available in chip form or in low-cost plastic packages, are fabricated with a low-noise indium-gallium-phosphate (InGaP) gallium-arsenide (GaAs) heterojunction-bipolar-transistor (HBT) process.

A simplified block diagram of one of these MMIC prescalers is shown in **Fig. 1.** In addition to the fixed digital frequency-divider network, which is implemented using flip flops, the prescaler includes amplifiers on the input and output sides. These amplifiers are configured with complementary pairs to enable either single-ended or differential-mode operation at the input and output ports. The die form of the prescalers includes three control terminals for the following functions: input disable, DC power down, and RF output-power level select.

The small size of the chip prescalers makes them useful for compact multichip-module (MCM) packaging configurations. On the other hand, the chip form requires more critical die attach and wire-bonding assembly processes, while the plastic-package devices are com-



1. This block diagram shows the control terminals for a high-frequency MMIC prescaler.

patible with conventional high-speed automatic surface-mount-technology (SMT) assembly equipment. Add to this the ruggedness of the plastic package MMIC prescaler, and it becomes the configuration of choice for use in many types of microwave assemblies that are intended for low-cost, high-volume applications. The HBT prescalers are designed for use with a single +5-VDC ±5-percent (±0.25-VDC) supply. The +5-VDC supply should be well-conditioned to minimize degradation of phase noise and be free from over-voltage transients that could damage the prescaler.

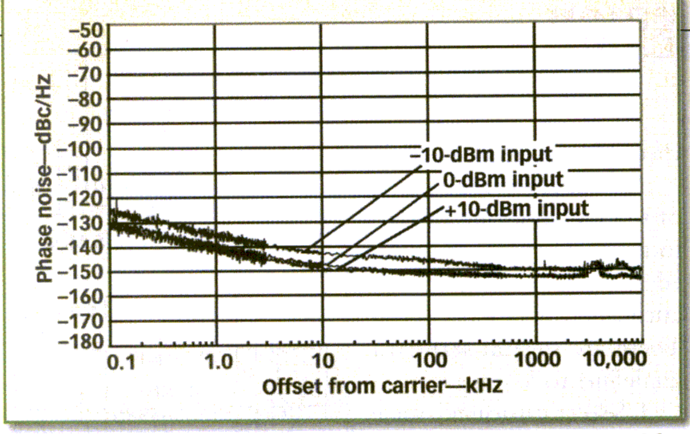
The +5-VDC prescalers include an internal monolithic 15-pF power-supply decoupling capacitor. However, a pair of external decoupling capacitors connected between V_{cc} and ground is highly recommended. To bypass lower frequencies, one capacitor should be relatively high in capacitance. A 1-to-10-μF tantalum chip or multilayer ceramic capacitor is recommended. The second capacitor should be relatively low in capacitance (such as 300 pF) to effectively bypass high-frequency components and to minimize prescaler input-to-output coupling.

The two decoupling capacitors, especially the lower capacitance one, should be placed as close as possible to the V_{cc} connection on the prescaler. The connection from the ground electrode of the capacitors to the circuit ground plane should be kept short. For chip prescalers, a single-layer 300-pF capacitor is recommended. The ground side of the single-layer capacitor should be attached to the carrier using conductive epoxy, or gold-tin (AuSn) eutectic solder, then the connection should be made to at least one of the V_{cc} pads on the chip using one or more wirebonds.

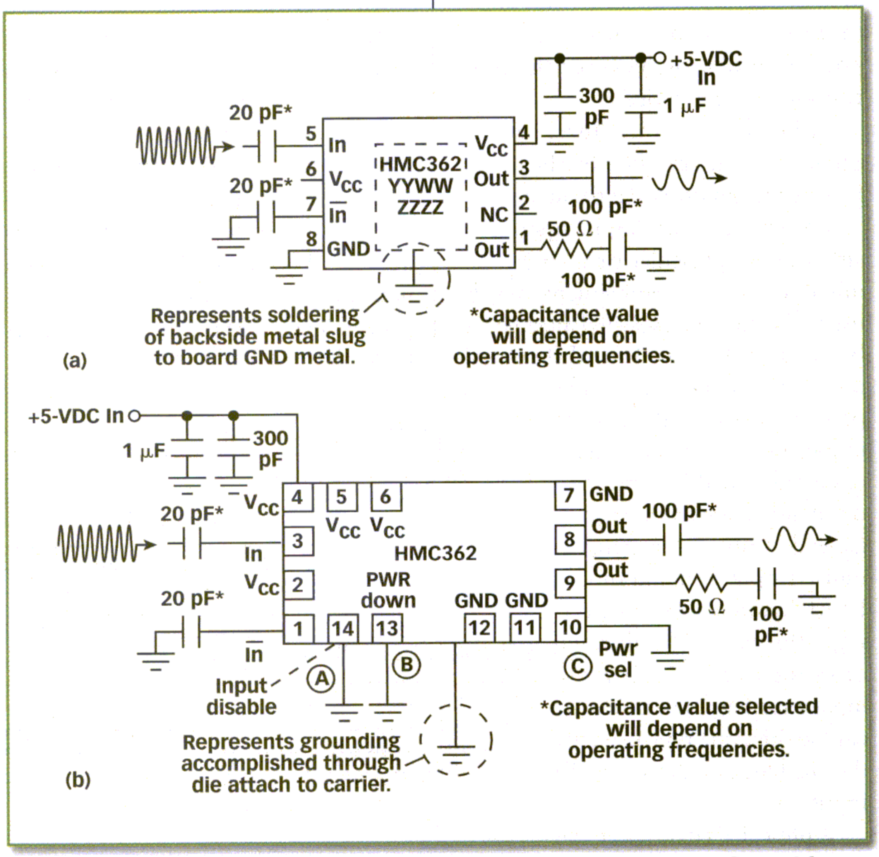
For normal prescaler operation, the power-down control terminal, present only on the bare die-form devices and not on plastic package parts, must be grounded. Applying +5 VDC to this pin puts the entire device into a power-down standby mode where the DC current consumption is reduced to a few mA. The DC power-down terminal can be controlled electronically with the output of a complementary-metal-oxide-semiconductor (CMOS) logic gate operating with a +5-VDC supply, or by connecting it to an open-collector logic device with a 1-kΩ pull-up resistor tied to the +5-VDC supply. The nominal source current required from the +5-VDC control signal to this input is less than 1 mA.

The nominal AC input impedance for each of the two differential prescaler inputs is a resistive 50 Ω referenced to ground. Although these are differential input ports, the prescaler can also operate as an unbalanced (single-ended) device.

Both prescaler input lines must be DC blocked from external circuitry. In the prescaler's internal circuitry, these inputs are DC coupled, with precisely controlled DC operating levels. Any external DC connection to the input lines can disturb these critical DC operating levels and degrade or disable operation of the device. If the input signal lines are already DC blocked by other circuitry at the source of the prescaler input signal, then DC blocking capacitors at the prescaler input can be omitted. This may be the case when the input lines are fed by microstrip coupled-line filters, high-pass filters, or from other circuitry that



2. The phase noise for a divide-by-four prescaler was measured for three different input-power levels.

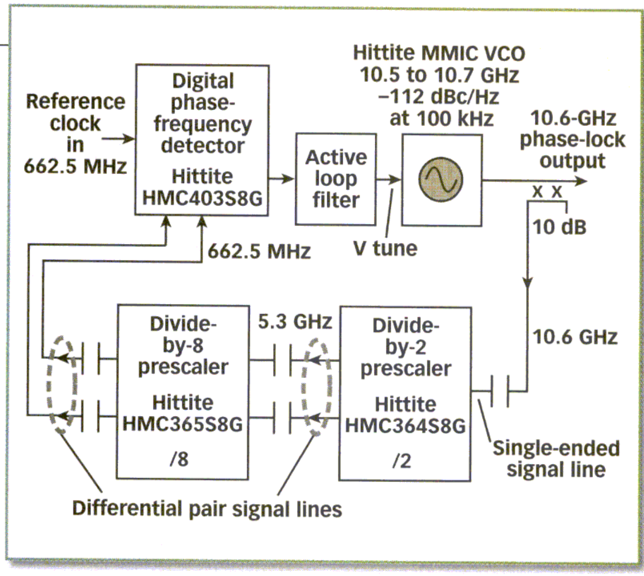


3. The external connections for this plastic-packaged prescaler are configured for single-ended operation (a). The chip version of the prescaler (b) is also configured for single-ended operation.

effectively provides a DC block.

The capacitance and style of the DC blocking capacitors should be properly selected to not cause significant attenuation of the input signal. There are several standard guidelines to follow:

1. Select a capacitance value which, for the lowest frequency of prescaler operation, presents a value of reactance which is small relative to 50 Ω, approximately 3 Ω or less. For example, if 10 GHz is the minimum input frequency, at least 5-pF DC blocking capacitance should be used. For a minimum input frequency of 1 GHz, at least 50-pF capacitance should be used.



4. This phase-locked 10.6-GHz oscillator was developed as a reference clock for SONET applications.

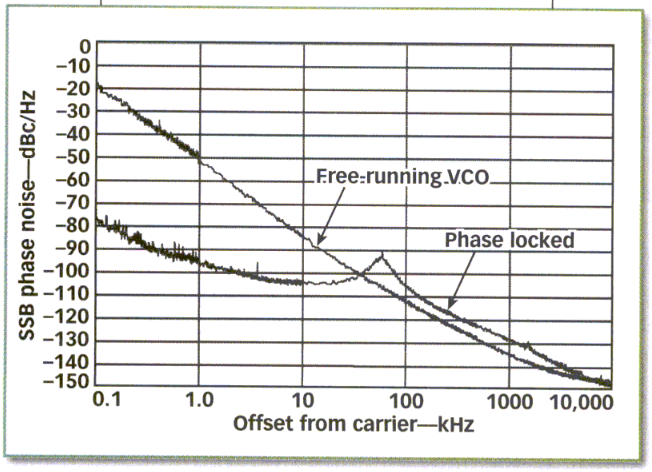
2. From the vendor's data for the capacitor, check that its self-resonant frequency (SRF) is higher than the highest input frequency of prescaler operation. For DC blocking, it may be acceptable to operate at or somewhat above the series SRF of a capacitor. To avoid potential problems, however, this practice should be avoided.

3. For best performance, select a capacitor style specified by the manufacturer for microwave frequencies.

For single-ended input operation, either of the prescaler's two input terminals (IN or IN Complement) may be used. The input line selected must be DC blocked as described for differential-mode input connections. The prescaler will operate with the unused input terminal left open circuited, but this is not recommended. Instead, for single-ended input operation, the unused input terminal should be AC grounded by connecting it through a DC blocking capacitor to ground to improve input sensitivity. The capacitor can be the same type used for the DC block on the active RF input line.

A prescaler's data sheet will provide a recommended operating window for input power. Typically, a device will operate over a greater than 25-dB range of input-power levels. However, toward the specified upper-frequency range of the device, the window is reduced in range to approximately 10 dB. For differential-mode input operation, the input-power range in the data sheet refers to total power from the two input lines, not the power on each line.

At the lower end of the recommended input-power operating window, the prescaler's output phase noise will begin to degrade. As an example, the output phase noise of a model HMC365 divide-by-four prescaler from Hittite Microwave Corp. was characterized for three different single-ended input-power levels at 6.65 GHz **(Fig. 2)**. The output phase noise first decreases for increasing input-power level. Then, above an input-power level of 0 dBm, the improvement in phase noise flattens out. Thus, for phase-noise-critical applications, it is better to operate this particular prescaler at input-power levels of 0 dBm or above, although input-power levels above the recommended maximum should be avoided. As the input power increases beyond the rated maximum level, spurious signals will begin to appear at the prescaler's output port. As the input power is further increased, at some level the device will become overstressed and device degradation may result.



5. The measured phase noise of the 10.6-GHz PLL oscillator is compared with the phase noise of the free-running (unlocked) 10.6-GHz VCO.

When possible, differential prescaler connections should always be made. To use only one side of an available differential input pair is effectively throwing away one-half of the input signal power, and input-power level may be at a premium. In addition, the differential mode of operation on the input side may provide some rejection of common-mode noise present on the input lines.

A MMIC prescaler offers a substantial amount of gain at its input and output ports. For conditions of no RF input signal or inadequate RF input signal level, the prescaler may go into a low-level uncontrolled oscillation. The general cause of this is feedback from device output to input in the presence of the high device gain. Prescaler self-oscillation can be avoided by providing an input signal of sufficient amplitude (as specified on a product data sheet).

For applications where an input signal is not continuously present, self-oscillation can be suppressed by active control of the input-disable function (available on the chip prescalers only). For normal operation, the input-disable pin must be grounded. (On plastic-packaged prescalers, the input-disable pin is permanently grounded internally.) When +5 VDC is applied to the chip input-disable pin, the output port of the prescaler's input amplifier is forced to a

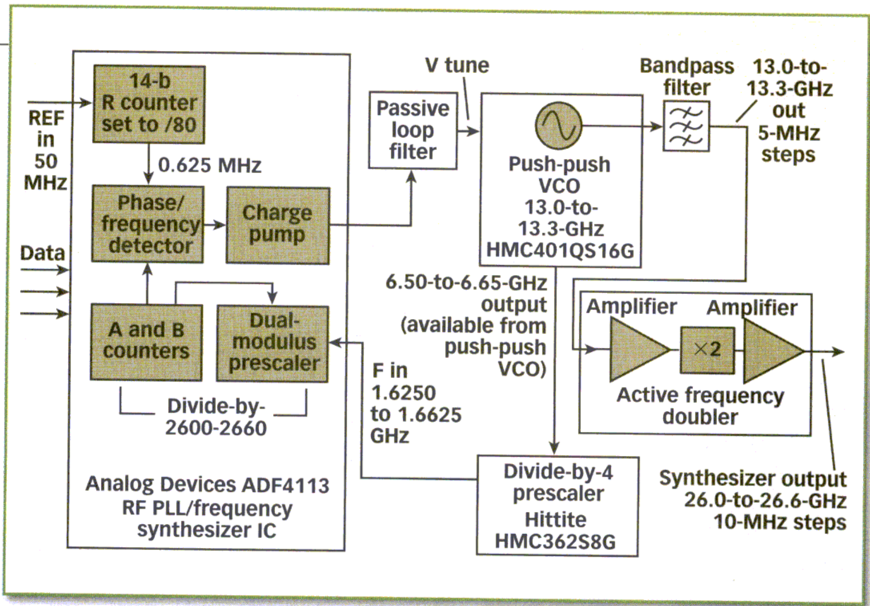
predetermined logic state, effectively placing the prescaler into a hold state, with one complementary output set at high and one set at low.

For high-speed operation, the input-disable pin can be driven by the output of a CMOS logic gate, or by connecting it to an open-collector output of a logic device with a 10-kΩ pull-up resistor tied to the +5-VDC supply. The nominal source that is current required from the +5-VDC control signal to drive the input-disable is less than 0.1 mA.

The output lines of the MMIC prescalers are also configured as a differential pair. The nominal RF output impedance of each output line is 50 Ω. Single-ended loads can be driven by simply using only one of the two output terminals. Being a digital device, the prescaler's output signal levels are independent of the input-power level, as long as the prescaler is used within its specified input signal range. There are some output variations as a function of frequency, however, as noted on each product data sheet (as specified for single-ended outputs). When properly configured to drive a differential load, the prescaler can deliver twice the amount of single-ended output power.

As with the input lines, the prescaler's output signal lines must be DC blocked. The same guidelines for selecting the input DC blocking capacitors can be applied to the choice of output DC blocking capacitors. However, the frequency at the output is lower than at the input and, consequently, the resulting minimum required capacitance value will be higher, by a factor equal to the division ratio.

For single-ended operation, there are two options for the unused output terminal. It can be left in an open-circuit condition, which is not recommended. Or a 50-Ω DC-blocked RF termination can be provided for this output terminal by connecting it to a 50-Ω resistor in series with a DC blocking capacitor, one side of which is grounded. The second approach helps reduce reverse-signal leakage, which is the appearance of the lower-frequency prescaler output signals (essentially



6. This 26.0-to-26.6-GHz frequency synthesizer was constructed with a divide-by-four prescaler and a commercial fractional-N synthesizer IC.

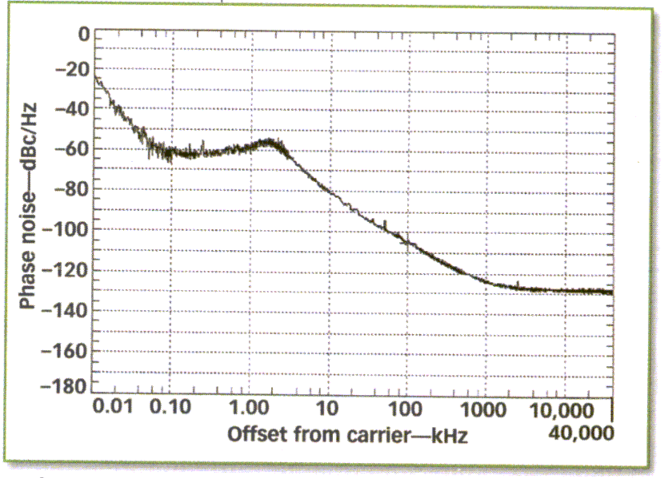
spurious signals) at the prescaler's higher-frequency input ports.

In some cases, this spurious distortion could propagate backward through a design and degrade signals in circuitry prior to the prescaler input terminals. Reverse leakage is normally characterized on a data sheet as a function of frequency for the two conditions of the unused output port: terminated and unterminated. With the output of the prescaler operating in differential mode into a load with 50 Ω per leg, the two output terminals are effectively terminated and a minimum reverse-leakage condition is achieved.

The output-power-select control terminal enables selection between two prescaler output-level operating modes. Direct DC grounding of the output-power-select terminal provides full RF output power, while leaving the output-power-select terminal open-circuited results in a reduced output-power mode, with approximately 6-dB reduction in output level compared to the full-level mode. In the reduced RF output-power mode, the DC current drawn by the prescaler drops by approximately 20 mA. The reduced output mode can be useful in those applications where DC power consumption is critical. The output-power-select line is generally not directly controllable by standard logic-gate outputs. Normally, this line will be hardwired to ground for full-power operation, or left open circuited for reduced output-power operation.

The packaged prescalers are supplied in an eight-lead small-outline-IC (SOIC) surface-mount housing. An exposed rectangular copper (Cu) alloy slug on the bottom side of the package is plated with tin-lead (SnPb) solder. The metal slug provides a good RF ground and is the primary heat-transfer interface for the MMIC.

Continued on page 132



7. The measured phase noise of the 26.0-to-26.6-GHz frequency synthesizer is shown for offset frequencies from 10 Hz to 40 MHz.

SPECIAL report

Specifying Microwave Voltage-Controlled Oscillators

Selecting a microwave voltage-tuned oscillator is a matter of fitting a desired set of performance specifications to the requirements of a particular application.

JACK BROWNE
Publisher/Editor

Frequency generation is a vital function in most electronic systems, and essential to wireless communications. In many modern communications systems, a voltage-controlled oscillator (VCO) is the signal source of choice, given its capability to tune in frequency rapidly in phase-locked loop (PLL) and other frequency-synthesized circuits. Selecting a VCO for a particular application is generally a matter of understanding the key specifications and weighing the importance of each performance parameter for that application.

A VCO can be thought of as an amplifier with high feedback from its output to its input, causing an unstable condition where oscillation occurs. The resonant frequency of the circuit containing the amplifier is the frequency of oscillation, and oscillation will continue provided that the amplifier can generate energy equal to the energy lost in the dissipative elements (such as capacitors and resistors) of the resonator. The active device (amplifier) in a VCO can be either a silicon (Si) bipolar transistor or a gallium-arsenide (GaAs) metal-semiconductor field-effect transistor (MESFET), with the latter usually employed in higher-frequency microwave and millimeter-wave designs.

By changing the capacitance in a VCO's resonant circuit, the frequency of oscillation will change. This can be achieved through the use of a voltage-dependent capacitive element, such as a varactor diode. The capacitance of the diode changes depending on the voltage applied to it, allowing the frequency of the VCO's resonant circuitry to be adjusted. Varactor diodes, with a relatively limited variable-capacitance range, are generally employed in narrowband VCOs. In contrast, hyperabrupt diodes, with a wide capacitance tuning range, are employed in VCOs that must tune over wider frequency ranges. The trade-off for a wider frequency-tuning range is generally some sacrifice in tuning linearity, which is the relationship of the amount of voltage

This compact VCO is typical of modern surface-mount devices. This particular model operates from 5220 to 5420 MHz in a package measuring only 0.50 × 0.50 × 0.13 in. (1.27 × 1.27 × 0.33 cm) (Photo courtesy of Z-Communications, San Diego, CA.)

applied to the change in frequency.

VCOs are available from a long list of suppliers in a variety of sizes and packages, including those with coaxial connectors, drop-in housings, and surface-mount packages. In selecting a VCO for a particular application, a designer is faced with some obvious choices concerning frequency range and desired size/package. Beyond making sure that an oscillator covers the required tuning range, it must also provide adequate output power, have minimal phase noise, suffer low levels of harmonic and spurious noise, provide good tuning linearity, and be resistant to variations in supply voltage and load.

In essence, these characteristics summarize the key specifications that define a VCO. For example, a VCO's output power, which is presented in terms of dBm, is also characterized in terms of flatness across the tuning range of the oscillator. So, an oscillator with rated output power of +7 dBm and output-power flatness of ±1 dB from 1900 to 2000 MHz, may actually provide as little as +6-dBm output power or up to +8-dBm output power at different frequencies across the tuning range. Both possibilities must be accounted for when preparing a system design. It should also be noted that since the output power of active devices will vary as a function of temperature, the output-power variations due to changes in temperature should be considered when selecting a VCO.

In terms of tuning range, a narrowband VCO may often be specified by a percentage of tunable bandwidth rather than a start-and-stop frequency. For example, a VCO with a center frequency of 1000 MHz and a 5-percent bandwidth has a total tuning range of 50 MHz, or start-and-stop frequencies of 975 to 1025 MHz. Wideband VCOs, which can offer tuning ranges of one octave or more, are generally specified in terms of their tuning range.

Phase noise is one measure of the frequency instability of an oscillator (and is often referred to as jitter in digital or optical terms). Phase noise in a VCO is contributed by the active devices, such as the bipolar transistor and the tuning diodes. It is measured in a 1-Hz bandwidth that is centered at various offsets from the carrier frequency, starting from 1 Hz and extending as far as 40 MHz from the carrier (the limit of most commercial test equipment). Phase noise is presented as a ratio of an oscillator's output power divided by its noise power, in units of dBc/Hz. A typical phase-noise specification might be −100 dBc/Hz at an offset of 10 kHz from the carrier. The phase noise of a VCO drops rapidly with increasing distance from the carrier, eventually reaching a constant noise level or noise floor, typically at offsets of 1 MHz or greater from the carrier.

Harmonic generation is a necessary evil of oscillator circuits. An oscillator with a desired fundamental output frequency of 1 GHz will also produce energy at the second-harmonic frequency of 2 GHz (as well as the third-harmonic frequency of 3 GHz, and so on). The level of these harmonic signals is generally specified as the amount of suppression (in decibels) or as the output level of the harmonics relative to the carrier output level (in dBc). A typical VCO harmonic value is −10 dBc. Similarly, spurious signals are nonharmonically related signal components that are typically caused by sources outside of the oscillator circuitry, such as power-line noise. As with harmonics, spurious noise is specified in terms of suppression (in decibels) or in terms of spurious output power that is relative to the level of the fundamental carrier (in dBc).

Tuning sensitivity and linearity are of particular importance to designers and users of PLLs and other frequency-synthesis circuitry. Sensitivity, which is usually specified in terms of megahertz/voltage, refers to the amount of change in oscillator frequency that occurs for a particular amount of applied tuning voltage. Tuning sensitivity can be as low as a few megahertz per volt or up to 50 MHz/V, depending upon the oscillator's tuning-range and the tuning voltage range. Tuning linearity refers to the predictability of a VCO's change in frequency as a function of tuning voltage. Usually the ideal plot of tuning linearity is a straight line, without breaks in the tuning range (where the oscillator stops operating with a certain tuning voltage).

Since a VCO must operate under conditions of changing supply voltage and load, it is specified in terms of the amount of frequency shift that will occur for a particular change in supply voltage (frequency pushing) and for the frequency shift that will occur for a particular load VSWR. Frequency pushing is specified in terms of megahertz/voltage, while pulling is given in terms of a particular amount of frequency deviation (such as 10 MHz or ±5 MHz) with reference to a given load condition (such as a VSWR of 2.0:1).

Recent trends in VCOs include the development of more compact surface-mount components for miniaturized communications systems, such as wireless handsets, base stations, digital radios, and satellite-communications terminals. As an example, the V940ME11 VCO from Z-Communications (San

Understanding VCO performance

Those in need of a well-written tutorial introduction to VCOs will find a seven-page application note, entitled "Voltage Controlled Oscillators," in the current *Designer's Handbook* from Synergy Microwave Corp. (Paterson, NJ, www.synergymwave.com). The article covers the types of oscillator circuits, such as Clapp oscillators, that are used in VCOs, as well as the basic equation for achieving oscillation, the capacitance-versus-voltage curves for a variety of different tuning diodes, a sample circuit for a wideband VCO, and a brief discussion on oscillator noise. The note includes details on a linear approach to the design of low-noise VCOs and provides a listing of key VCO terms, such as harmonics, spurious, and phase noise.

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VAT-6	6	0.10	1.15
VAT-7	7	0.10	1.15
VAT-8	8	0.10	1.20
VAT-9	9	0.10	1.15
VAT-10	10	0.20	1.20
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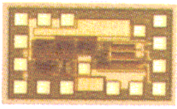
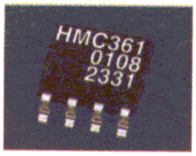
		DC - 12.5	-145 dBc/Hz	HMC364S8G
÷4	High Efficiency Med. Output Power	DC - 12.0	-149 dBc/Hz	HMC362
		DC - 12.0	-149 dBc/Hz	HMC362S8G
÷4	High Frequency High Output Power	DC - 13.0	-151 dBc/Hz	HMC365
		DC - 12.5	-151 dBc/Hz	HMC365S8G
÷8	High Efficiency Med. Output Power	DC - 12.0	-153 dBc/Hz	HMC363
		DC - 12.0	-153 dBc/Hz	HMC363S8G

Divide-by-2

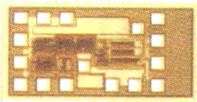


HMC361

HMC361S8G



Divide-by-2



HMC364

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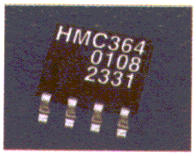
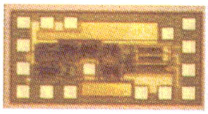
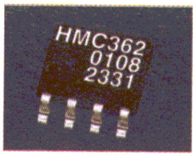


Divide-by-4

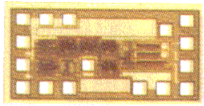


HMC362

HMC362S8G

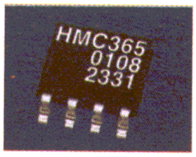


Divide-by-4

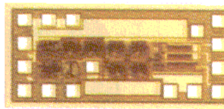


HMC365

HMC365S8G

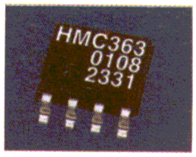


Divide-by-8



HMC363

HMC363S8G



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Diego, CA) is designed for Unlicensed National Information Infrastructure (UNII) band applications, such as HIPERLAN wireless local-area-network (WLAN) systems, from 5220 to 5420 MHz **(see figure)**. Its surface-mount package measures only 0.50 × 0.50 × 0.13 in. (1.27 × 12.7 × 0.33 cm). The oscillator requires only +0.5 to +4.5 VDC to control its 200-MHz tuning range.

An extensive line of miniature surface-mount VCOs is available from Synergy Microwave Corp. (Paterson, NJ), encased in the company's 174 package measuring only 0.50 × 0.50 × 0.22 in. (1.27 × 1.27 × 0.56 cm). The line includes models operating from 40 to 100 MHz through models up to 6100 MHz. The company **(see sidebar)** also offers VCOs in a variety of other surface-mount and plug-in packages, as well as voltage-tuned oscillators based on ceramic coaxial resonators and surface-acoustic-wave (SAW) resonators for optical-communications applications (see *Microwaves & RF*, April 2001, p. 115).

Mini-Circuits (Brooklyn, NY) includes a line of compact VCOs in its extensive product catalog. Model JTOS-2200PA, which is available in a compact surface-mount package, is a new addition designed for applications from 2000 to 2200 MHz. It provides +6-dBm typical output power over that range, with −22-dBc typical harmonic suppression. The phase noise is −92 dBc/Hz offset 10 kHz from the carrier and −112 dBc/Hz offset 100 kHz from the carrier.

Vari-L (Denver, CO) also offers VCOs in its tiny surface-mount T package, which measures only 0.50 × 0.50 × 0.18 in. (1.27 × 12.7 × 0.46 cm). The company's model VCO190-1925T, for example, operates from 1650 to 2200 MHz with a typical tuning sensitivity of 60 MHz/V and typical output power of 0 dBm. The typical phase noise is −90 dBc/Hz offset 10 kHz from the carrier, dropping to −110 dBc/Hz offset 100 kHz from the carrier. Suppression of second harmonics is typically −10 dBc while spurious suppression is −90 dBc.

Amplifonix (Philadelphia, PA) offers lines of VCOs in a variety of compact housings, including TO-8, dual-inline-package (DIP), flat-pack, connectorized, and surface-mount packages. Models cover tuning ranges of 60 to 70 MHz through 2000 to 2500 MHz, with typical output levels exceeding +10 dBm. The maximum phase noise for most models is −100 dBc/Hz offset 100 kHz from the carrier, with typical harmonic suppression as good as −40 dBc for some models and typically exceeding −12 dBc for all models.

The VS series of surface-mount VCOs from MODCO, Inc. (Sparks, NV) is designed for high-density radio applications, such as in Personal Computer Memory Card International Association (PCMCIA) modem cards. The compact VCOs, which measure only 0.175 × 0.175 × 0.062 in. (0.445 × 0.455 × 0.157 cm), are available with up to 10-percent bandwidths from 700 to 5800 MHz. For a 3-GHz model, the phase noise is less than −105 dBc/Hz offset 100 kHz from the carrier.

IBM (Hopewell Junction, NY) entered the market for high-frequency VCOs several years ago based on the strength of its Si-germanium (SiGe) bipolar-complementary-metal-oxide-semiconductor (BiCMOS) process and with an order from a major wireless-handset manufacturer. The company now offers fully integrated VCOs with on-chip tank circuits and core VCOs (cVCOs) that require off-chip tank circuits. The latest addition to the IBM product line is a triband cVCO for Global System for Mobile Communications (GSM) applications. With an operating frequency range of 3.6 to 4.0 GHz, the VCO provides −3-dBm output power with measured phase noise of −147 dBc/Hz offset 3 MHz from the carrier and −162 dBc/Hz offset 20 MHz from the carrier.

Agilent Technologies (Santa Clara, CA) offers a line of VCOs in compact TO-8 housings. The company's VTO-8000 line is based on varactor tuning, while the VTO-9000 series uses hyperabrupt diode tuning. From the latter series, for example, the model VTO-9090 tunes from 900 to 16000 MHz with +10-dBm output power and phase noise of −100 dBc/Hz offset 50 kHz from the carrier. The firm recently announced the model VTO-0995-T with a center frequency of 9953.28 MHz specifically for OC-192 10-Gb/s optical-communications systems. The TO-8 oscillator is also available in a surface-mount version as model VTO-0995-S. The company has also announced the development of a family of surface-mount VCOs that is suitable for 40-Gb/s optical systems, including 19.906-GHz and 39.813 models with differential outputs and time jitter of better than 50 fs.

Seven years ago, Micronetics (Hudson, NH) acquired the M3500 series of VCOs from cellular-communications supplier Qualcomm (San Diego, CA). The company has since improved the oscillators, which are available in ranges from 100 to 3500 MHz, to include several models in 0.5 × 0.5-in. (1.27 × 1.27-cm) surface-mount packages. The oscillators typically provide more than +12-dBm output power with phase noise of less than −94 dBc/Hz offset 10 kHz from the carrier. MRF

PCB layout is key to MMIC performance

THE HIGHER the operating frequency of a MMIC is, the greater the care that is required to design and layout its PCB. With today's devices sometimes running at tens of gigahertz, board layout and the materials used must be given close scrutiny according to an application note from Hittite Microwave Corp.

Entitled "Layout Guidelines for MMIC Components," the 3-page note was written to guide designers on board layout for two new MMICs from the company—a voltage variable attenuator and a mixer. But the layout principles could be useful to all designers of gigahertz frequency-range communications equipment.

The attenuator and mixer are supplied in ceramic and plastic surface-mount packages intended to operate from DC to 15 GHz. The note suggests that to obtain the best test data, the devices should use a transmission line that is a grounded CPWG on Rogers 4350 material or equivalent. CPWG is used instead of the more common microstrip since it permits a 30-percent reduction in line width for 50-Ω lines. When fabricated on similar substrate materials and at a frequency of 10 GHz, CPWG supports a larger ground plane between RF lines thus enhancing isolation and reducing leakage.

CPWG fabrication requires via holes to connect the top and bottom ground planes along the RF transmission line.The placement of these holes is critical to the board's performance, as is the plating of the holes. An evaluation board consists of two layers of the 4350 material bonded together with 4403 adhesive to provide an overall thickness of 0.062 to 0.067 in. (0.0157 to 0.0170 mm). The 4350 material is Cu clad on both sides, resulting in four metal layers. Plating of the external metal layers must be carefully controlled to maintain the correct impedance of the RF lines.

Careful board fabrication is only part of the process in obtaining accurate data when evaluating a MMIC. Hittite also uses an evaluation board calibrated to a custom set of PCB standards. The idea is to be able to isolate the performance of the DUT from the losses and mismatches of the board. The measurement accuracy depends on the standards and the transitions from connector to PCB

The note can be downloaded from the company's website.

Hittite Microwave Corp., 12 Elizabeth Dr., Chelmsford, MA 01824; (978) 250-3343, FAX: (978) 250-3373, Internet: www.hittite.com.

Enter No. 194 at www.mwrf.com

Real-world design tips reduce RF noise

AT RADIO frequencies and above, basic circuit elements—resistors, capacitors, inductors, wires—take on an entirely different nature than at lower frequencies. The details of their operation are covered in an informal but comprehensive way in the application note, "Practical R.F. Design." This note complements the system-level view of high-frequency interference that was presented in this column last month (see "Stick To The Basics For Low-Noise Design," *Microwaves & RF*, August 2001, p. 183.)

The note begins by showing why the "simple" resistor, capacitor, and inductor are not simple at all at high frequencies. Each of these component's parasitic elements must be recognized and accounted for at high frequency (i.e., capacitance in resistors, resistance in capacitors, etc.). A good discussion of the potential problems of inductors at high frequency is provided. Also involved here is the inductance of conductors, which while low, can result in impedances that can cause unforseen problems at high frequencies.

As in last month's application note, the methods and types of grounding have a major influence on minimizing noise. The note claims that the best ground arrangement for a PCB is a ground plane since it offers minimal resistance and inductance, but there could be circumstances where high currents and high frequencies combine to create significant voltage drops across the plane.

Decoupling is an issue that causes problems at high frequency. All ICs on a board must be individually decoupled with low-inductance capacitors having short leads and PC tracks. The note claims that "even 2-3 mm of extra lead/track length may make the difference between success and failure of a circuit layout."

To sum up the key points, a successful RF design will pay close attention to resistance, capacitance, inductance, grounding, and decoupling, and will separate sensitive circuits from noisy ones. The note can be downloaded from the manufacturer's website.

Analog Devices, Inc. Three Technology Way, Norwood, MA 02062; (781) 329-4700, Internet: www.analog.com.

Enter No. 195 at www.mwrf.com

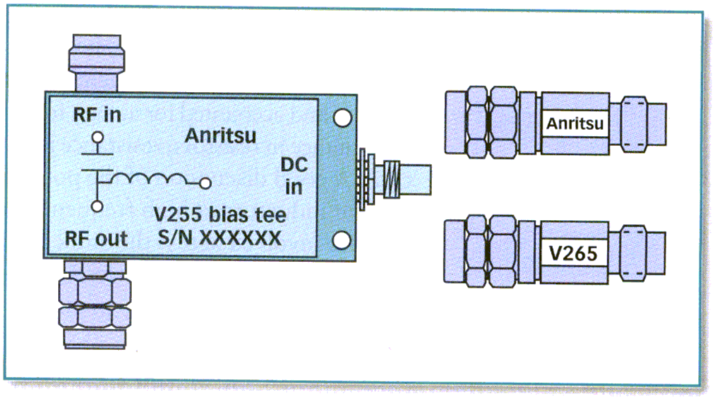
Bias Tee And O Illuminate

Optical-communications systems promise the wide bandwidth needed for instant Internet access and a wide range of voice, data, and multimedia services to the home and office. But despite the wide-bandwidth potential of lightwave systems, they still require electrical components and connectors working in conjunction with the glass fibers and photodetectors. Two of the electrical components developed by Anritsu Co. (Morgan Hill, CA) to support these applications, a bias tee and a DC block, incorporate a newly refined configured high-frequency coaxial connector to provide near DC-to-65-GHz bandwidths to support high-speed optical- and data-communications systems.

RAVINDER GILL
Product Marketing Manager
North American Measurements Group, Anritsu Co., 1155 East Collins Blvd., Richardson, TX 75081; (800) 267-4878, FAX: (972) 671-1877, Internet: www.us.anritsu.com.

To meet the electrical requirements for passive components employed in a high-speed optical-communications network, Anritsu Co. has developed the model V255 bias tee and the model V265 DC block. The components are armed with the company's high-frequency V Connector®, which has already been accepted as a standard coaxial interface for OC-768 optical-communications applications due to the connector's excellent impedance match across such a wide bandwidth (65 GHz).

In both components, the input ports have female V Connectors, the



1. The model V255 bias tee and model V265 DC block rely on the low-loss performance of the V Connector for operation through 65 GHz.

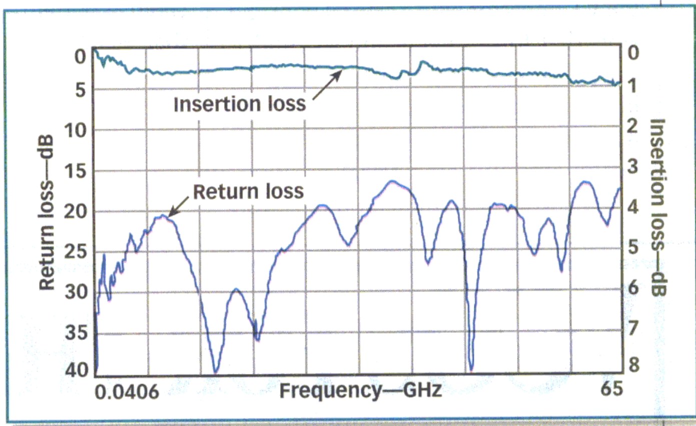
DC Block 65 GHz

These two high-frequency components leverage the performance of a precision coaxial connector to deliver low-loss signals through 65 GHz.

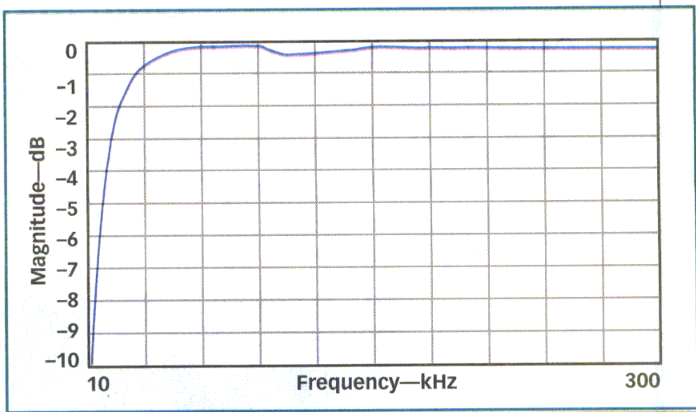
output ports have male V Connectors, and Gen II Bias T DC ports are equipped with an SMC connector **(Fig. 1)**.

The integrated model V255 bias tee and model V265 DC block provide excellent performance from 50 kHz to 65 GHz at operating temperatures from 0 to +80°C. The maximum insertion loss for the V255 bias tee is 1.5 dB, with minimum return loss of −12 dB over the full operating temperature range. The insertion-loss and return-loss performance of the model V255 bias tee were characterized at room temperature from 40 MHz to 65 GHz, revealing that actual performance far exceeds the published specifications **(Fig. 2)**. Measurements were taken with a Lightning 37397C series vector network analyzer (VNA) from Anritsu Co. The applied bias current during the measurements was 400 mA.

The low-frequency insertion-loss performance of the bias tee is shown in **Fig. 3**. The data were taken with a model MS4630B scalar network analyzer (SNA) from Anritsu Co. The model V255 bias tee, which can simultaneously apply DC and RF drive signals to a device through a single input port, is designed to handle up to 400-mA DC bias current and +10-VDC DC bias voltage. Due to its high current-carrying capacity with such low losses and broadband frequency response, the model V255 bias tee is suitable for use in biasing 40-Gb/s optical modulators and 40-Gb/s data drivers.



2. The high-frequency insertion-loss and return-loss performance levels of the model V255 bias tee were evaluated with a Lightning 360 series VNA.



3. The low-frequency insertion-loss performance of the model V265 bias tee was measured with a model MS4630B scalar network analyzer.

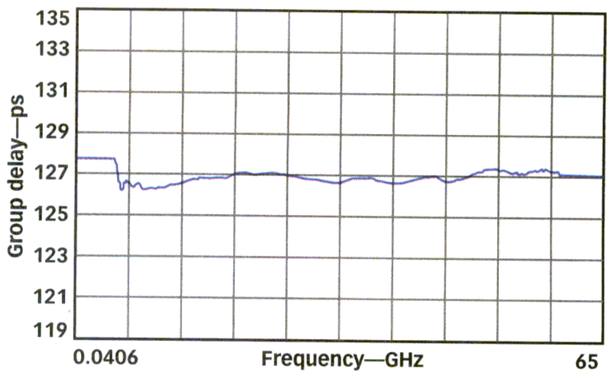
At data rates as fast as 40 Gb/s, the insertion loss through a device is critical. A bias tee used to bias an optical modulator must have low rise times and flat group-delay response or it will distort the signal, adversely affecting the data transmitted through it and resulting in an increase in system bit-error rate (BER). With this in mind, the model V255 bias tee has been optimized for 40-Gb/s optical communications and other high-speed pulsed, data, or microwave applications. The V255 bias tee features fast rise times of better than 3 ps and flat group delay of better than ±1 ps.

Figure 4 shows the group-delay performance of the model V255 bias tee. The reference is set at 127 ps with a scale of 2 ps/div. Due to the superior performance of the V255 bias tee, it is virtually transparent within the circuit, supporting extremely accurate measurements within laboratory environments.

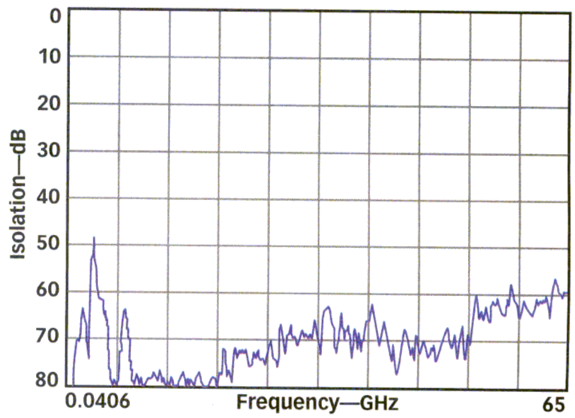
Figure 5 shows the typical isolation performance of the model V255 bias tee between its DC and RF ports. The model V255 second-generation (Gen II) bias tee exhibits better than 50-dB isolation across its full frequency range.

The high performance of the model V255 bias tee is achieved through Anritsu's patented true coaxial structure. The coaxial structure uses the V Connector's center conductor to provide an axially resilient coaxial connection.[1] This connection is comprised of a cylindrical center-conducting member of the V Connector with a central bore and slots that form fingers and a cylindrical pressure-contact member that is inserted into the cylindrical conducting medium. Both center-conductor member fingers and cylindrical pressure-

Continued on page 133



4. The group-delay performance of the model V255 bias tee was evaluated with a Lightning 360 series VNA. The reference was set at 127 ps with a scale of 2 ps/div.



5. The isolation between the bias tee's RF and DC ports was measured at better than 50 dB.

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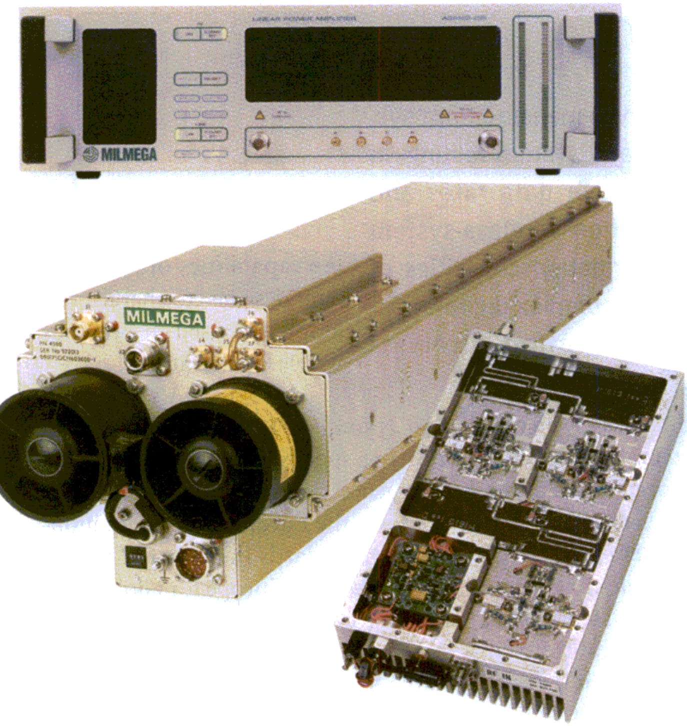
Milmega's strength lies in our ability to produce a family of high quality octave band solid state microwave amplifiers, and yet maintain the flexibility to design and build high quality amplifiers/amplifier systems to meet your specific requirements. This strength has created Milmega's global reputation as experts in the field of high power microwave engineering.



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Providing wide frequency ranges and producing field strengths over 200V/m.

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Operating over wide temperature ranges in rugged environments, Milmega amplifiers are built into sophisticated jamming systems.

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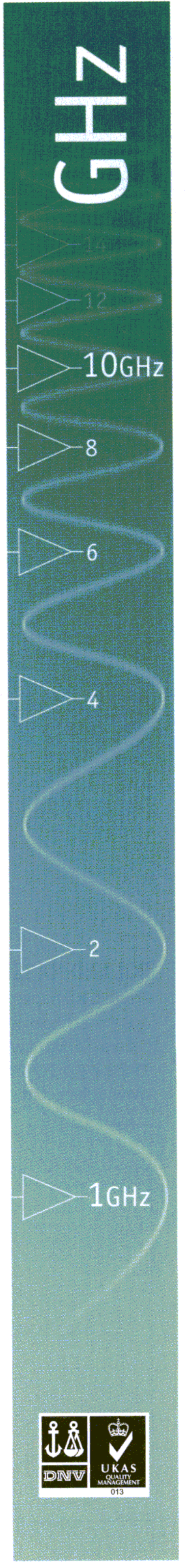

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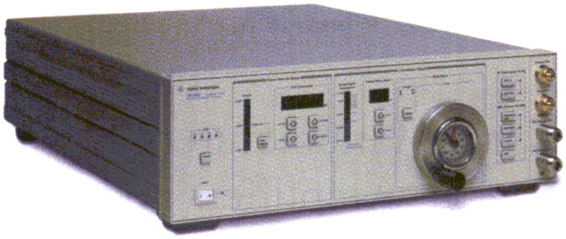
System Boosts Amplifier Test-Set Dynamic Range

This system can effectively extend measurement dynamic range by 25 dB or more when evaluating the performance of amplifiers designed for 2.5G and 3G wireless systems.

ROLAND HASSUN
Product Manager
Agilent Technologies, Inc., 1400 Fountaingrove Pkwy., Santa Rosa, CA 95403; (707) 577-2000, Internet: www.agilent.com.

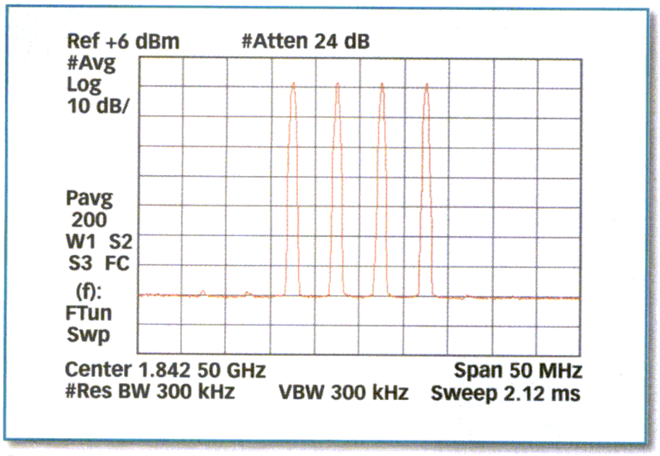
Dynamic range is a valuable commodity in measurement systems aimed at wireless component and system testing. In response to customer needs for higher data rates and mobile Internet access, next-generation wireless-system designers continue to push the information capacities of limited bandwidths using increasingly complex digital-modulation formats. These latest-generation wireless-system requirements challenge the dynamic-range capabilities of even the best spectrum analyzers. Fortunately, an enhancement for spectrum-analyzer-based systems has been developed by Agilent Technologies (Santa Rosa, CA) that overcomes the dynamic-range limitations of conventional systems. The N4256A amplifier distortion test set **(Fig. 1)** provides a dynamic-range improvement of up to 25 dB **(see table)**. It can measure spurious levels lower than −80 dBc in a 100-kHz bandwidth from 0.5 to 4.0 GHz, which is beyond the capabilities of most systems designed to characterize amplifiers used in previous wireless generations.



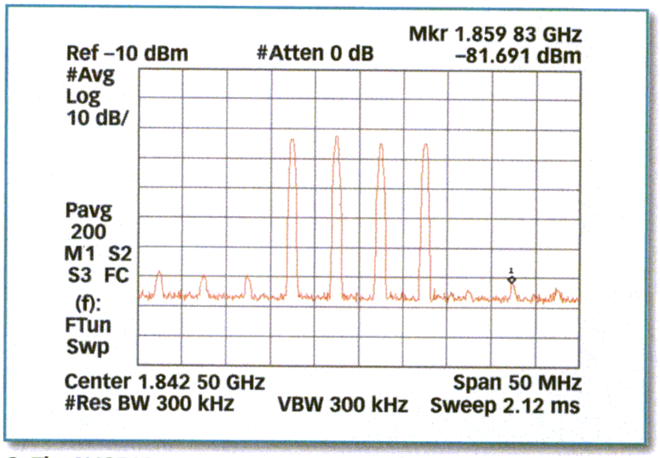
1. The N4256A measurement system provides approximately 25 dB more dynamic range than conventional distortion-measurement systems for evaluation of linear PAs.

The N4256A's contribution is underscored by the rigors of meeting specifications contained in next-generation wireless standards. For example, the European Telecommunications Standards Institute (ETSI) has set standards in Europe for Global System for Mobile Communications (GSM) systems based on the Enhanced Data Rates for Global Evolution (EDGE), and systems using wideband-code-division-multiple-access (WCDMA) techniques.

EDGE upgrades GSM systems from Gaussian minimum-shift-keying (GMSK) modulation to the more bandwidth-efficient eight-state phase-shift-keying (8PSK) modulation. The use of 8PSK compared to GMSK increases the peak-to-average power ratio (crest factor) of a power amplifier (PA) used in a base transceiver station (BTS) and places much greater demands on the linearity of the amplifier. For offset distances greater than 6 MHz from the carrier frequency, the spurious-emission specification is an extremely demanding −70



2. A spectrum analyzer was used to evaluate an MCPA driving four +39.3-dBm GSM signals with channel separation of 5 MHz. No apparent spurious content is visible.



3. The N4256A measurement system shows the previously invisible spurious content that is generated with the amplification of the four 5-MHz GSM channels in Fig. 2.

dBc or less in a 100-kHz bandwidth.

In a typical test system designed to measure intermodulation distortion (IMD) and adjacent-channel power ratio (ACPR), the device under test (DUT) is stimulated by a source supplying the appropriate signal [continuous wave (CW) for IMD or modulated single or multichannel carrier for ACPR]. The output of the DUT is connected to a high-power load, with some of the output signal coupled to a spectrum analyzer. Ideally, the source would produce a perfect stimulus spectrum with power only in the desired channel. In practice, the signal includes distortion sidebands that mask the performance of the DUT. The spectrum analyzer must determine the ratio of the DUT-generated distortion sidebands present in the adjacent and alternate channels to the power in the desired channel. With the latest multicarrier amplifiers, the difference in power between the two exceeds the dynamic range available in commercial spectrum analyzers.

These two problems—distortion present in the source signal and limited spectrum-analyzer dynamic range—effectively prevent accurate characterization of linear PAs. The N4256A was developed to solve theses problems. For a typical IMD or ACPR measurement setup, the N4256A is inserted between the source and spectrum analyzer, parallel with the DUT. The stimulus signal is sampled before the DUT, conditioned, and injected back into the path that is before the spectrum analyzer. Conditioning includes amplitude and phase matching.

Since the conditioned signal (the source stimulus) is 180 deg. out of phase with the DUT output signal, the distortion products that were generated within the source are highly cancelled, leaving only the distortion products generated within the DUT for measurement. In this cancellation process, the carrier at the DUT output is also cancelled to a large degree, reducing the dynamic range of the signal presented to the spectrum analyzer. Due to the lower power of this cancelled input signal, the spectrum analyzer's input attenuation can be safely reduced, and its preamplifier (if so equipped) can be turned on, greatly enhancing noise figure.

By effectively canceling distortion products that are generated in the measurement system's source and reducing system noise figure as described earlier, the N4256A is able to dramatically increase the effective dynamic range of distortion measurements. It has demonstrated effective dynamic-range enhancement in tests at Powerwave Technologies (Irvine, CA) of up to 30 dB.

The N4256A system was used to evaluate several different amplifiers, including a GSM multichannel PA (MCPA), and compared to conventional spectrum-analyzer measurements. The GSM amplifier was used to generate four independent +39.3-dBm modulated output signals with channel separation of 5 MHz. The total RF power of the four combined signals was +45.3 dBm, or 33.9 W. When displayed on a spectrum analyzer alone **(Fig. 2)**, the four carriers are visible, but little or no spurious content is discernible. When the N4256A system is used for the same measurement **(Fig. 3)**, IMD is apparent above the noise floor on the analyzer's screen. The extra dynamic range that is provided by the N4256A system (up to 25 dB) made it possible to detect and analyze these spurious signals. P&A: $31,000; 6 wks. **Agilent Technologies, Inc., Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; Internet: www.agilent.com.**

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The N4256A at a glance

FREQUENCY RANGE	0.5 TO 4.0 GHz
Input power	1 W maximum (50-W input power)
VSWR	
DC to 2 GHz	1.20:1
2 to 4 GHz	1.30:1
Isolation	90 dB minimum
Dynamic-range extension	25 dB typical
Power consumption	55 W at +115 VAC

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Will RF Coexistence Erode Performance?

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Will Bluetooth Ever be Cost-Effective?

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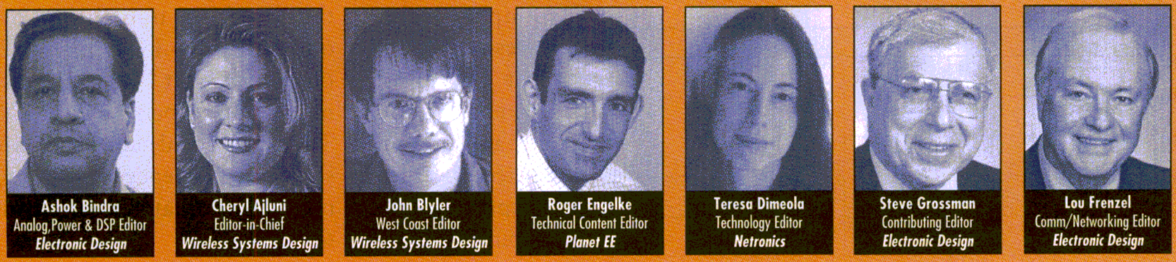

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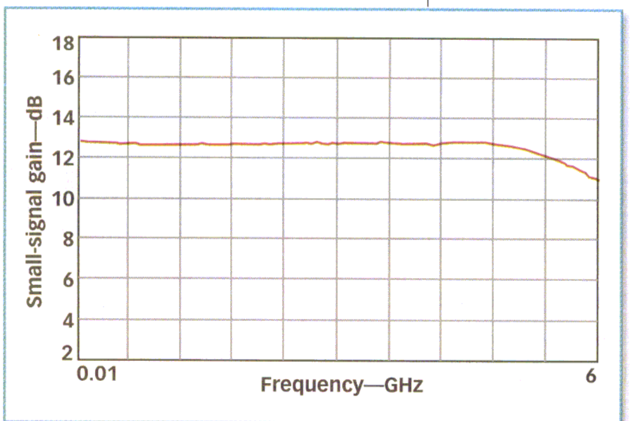
Linear HBT Amplifiers Arrive From New Source

A company that was not previously associated with amplifiers has entered the market for high-gain, high-linearity products with a pair of low-cost InGaP-based HBT amplifiers.

JACK BROWNE
Publisher/Editor

Linearity requirements for wireless communications systems continue to push the demand for high-performance heterojunction-bipolar-transistor (HBT) amplifiers. Numerous suppliers exist, although a new supplier—Metelics Corp. (Sunnyvale, CA)—has joined the market with several indium-gallium-phosphide (InGaP)-based HBT amplifiers that combine high linearity with high efficiency. The first two products, models MMA701 and MMA710, offer operating bandwidths of 4 and 5 GHz, respectively.

The company, which is well-known for its semiconductors and diode-based components, enters a mature market with extremely linear amplifiers. The MMA701, for example, which is designed for use from 100 MHz to 4 GHz, features an output third-order intercept point (IP3) of +45 dBm at 2.1 GHz. The amplifier also achieves +25-dBm output power at 1-dB compression. In linearity applications, the difference between the output IP3 and the output 1-dB compression point is often used as a figure of merit. For this amplifier, the 20-dB difference can be considered as good or better than most HBT amplifiers on the market. The MMA701 can be supplied in die form or in packages, including SOT-89 surface-mount housings. The amplifier achieves 12.2-dB typical gain with a single +5-VDC supply (and 130-mA current), with typical noise figure of 3 dB. The power-added efficiency (PAE) is typically 50 percent.



The MA710 Darlington HBT amplifier provides typically 12.5-dB gain across its operating range of 10 kHz to 5 GHz.

Designed for applications that range from 100 kHz to 5 GHz, the MMA710 is an HBT Darlington amplifier that is supplied in an SOT-89 surface-mount housing. Although it does not achieve the impressive linearity of the MMA701, the MMA710 nonetheless offers an output IP3 of +35-dBm and +21-dBm output power at 1-dB compression, for measurements performed at 500 MHz. The small-signal gain is 12.5 dB at that same test frequency **(see figure)**, with a noise figure that typically consists of 6.5 dB.

The packaged amplifiers can be supplied in bulk or tape-and-reel formats. **Metelics Corp., 975 Stewart Ave., Sunnyvale, CA 94086; (408) 737-8181, FAX: (408) 733-7645, e-mail: sales@metelics.com, Internet: www.metelics.com.** MRF

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Adapter Makes Blindmate Connections To 40 GHz

This SMP socket-to-socket adapter provides the RF performance and misalignment forgiveness needed for a variety of high-density microwave circuit and systems applications.

JACK BROWNE
Publisher/Editor

Blind-mate connectors came into prominence in the 1980s as practical solutions for interconnecting components and circuits in high-density enclosures. The SMP bullet adapter from W.L. Gore & Associates (Newark, DE) is designed to work with high-performance blind-mate connectors to provide flexible connectivity solutions through 40 GHz. The 50-Ω socket-to-socket adapter (see figure) simplifies the use of miniature push-on coaxial connectors such as the SMP and mini-SMP connectors used to interconnect modules and components within tightly packaged PCBs. The adapter tolerates significant misalignent in the radial and axial directions. In the radial direction, the adapter can forgive misalignment as poor as 0.020 in. (0.5 mm), while in the axial direction, the adapters will mate successfully even with up to 0.010-in. (9.3-mm) misalignment.

The SMP socket-to-socket adapter supports a line of blind-mate connectors for high-density packaging applications.

The SMP "bullet" adapter features BeCu center contacts and housing and PTFE dielectric. The housings and contacts feature Au plating on top of Ni plating to minimize reflections and loss. The SMP bullet adapter is rated for typical VSWR of less than 1.10:1 from DC to 23 GHz, less than 1.15:1 from 23 to 26.5 GHz, and less than 1.25:1 from 26.5 to 40 GHz. The RF leakage is only −80 dB at 3 GHz and −65 dB from 3 to 26.5 GHz.

The adapter is rated for a corona level of at least +190-VDC RMS at 70,000 ft., and dielectric withstanding voltage of +125-VDC RMS at the same altitude. The dielectric withstanding voltage improves to +500-VDC RMS at sea level. The adapter is characterized for an RF high potential of +325-VDC RMS at 5 MHz and minimum insulation resistance of 5000 MΩ. The maximum contact resistance for the center conductor is 6 mΩ, while the maximum contact resistance for the outer conductor is 2 mΩ. **W.L. Gore & Associates, Inc., 750 Ott's Chapel Rd., Newark, DE 19714; (800) 445-4673, (302) 292-5100, Internet: www.gore.com/electronics.**

Enter No. 44 at www.mwrf.com

Jitter Analyzers Help Solve Timing Problems

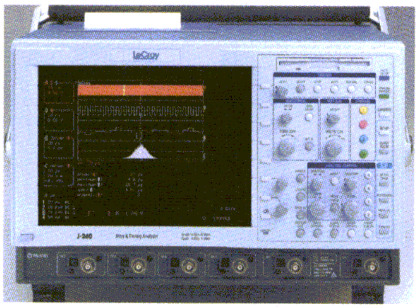
A line of jitter and timing analyzers performs measurements with 1-ps accuracy and provides analysis in the time, frequency, and statistical domains.

DON KELLER
Senior Editor

Timing is everything—especially in the world of modern electronic communications, where frequencies are measured in gigahertz and time is measured in nanoseconds and picoseconds. Measuring and analyzing the timing characteristics of communications sytems is critical to assessing their performance and troubleshooting problems. Jitter- and timing-analysis tools have been available as options on sophisticated oscilloscopes for some time. But, acknowledging the critical importance of these tools, test-equipment manufacturers such as LeCroy (Chestnut Ridge, NY) are now offering dedicated jitter and timing analyzers. The company recently introduced two of these instruments—models J250 and J260 **(see figure)**.

The two models are very similar in appearance and function. The main difference between them is that the model J260 has a 2-GHz bandwidth, while the J250's bandwidth is 1 GHz.

Both models have four input channels and a color screen that can display all four channels simultaneously. It is clear that these instruments are custom designed for jitter testing. They can display simultaneous clock, timing, statistical, and frequency-domain views. For example, they can display views of the raw signal, graphs of timing variations, statistical views of variations in clock period, frequency, cycle-to-cycle variation, pulse width, duty cycle, setup/hold jitter, and other important timing characteristics.



The J250 and J260 can show four signal characteristics simultaneously.

The analyzers are designed to simplify measurements that are difficult or impossible to make with non-dedicated instruments. Three examples of these measurements are accumulated jitter, half-cycle jitter, and PLL loop bandwidth (jitter-transfer function).

The analyzers can sample signals at rates up to 16 GSamples/s and can capture and record up to 32 Mpoints of data. Maximum jitter accuracy is 1 ps (with a 3-sigma confidence level in the J260). **LeCroy, 700 Chestnut Ridge Rd., Chestnut Ridge, NY 10977; (800) 453-2769, Internet: www.lecroy.com.**

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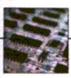

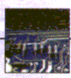

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Improper connection from the metal ground slug to a printed-circuit board (PCB) can result in reduced bandwidth performance as well as device overheating. The metal slug on the package must be soldered to a rectangular metal pad on the circuit board. The pad must be a good RF ground and a good heat sink. The size of the rectangular metal pad should be approximately the same size as the exposed metal slug on the package. For good heat sinking, the ground pad on the circuit board should include plated metal via holes extending to another large area of grounded metal on the board, such as a ground plane.

Soldering of this ground slug of the plastic package to the circuit board is best accomplished using solder paste and a solder-reflow process. Since the exposed surface of the metal slug is coated with SnPb solder, electrically conductive epoxy (which is not compatible with SnPb solder) must not be used for the attachment. A soldering iron is not suitable for this attachment process since the surfaces to be bonded are not exposed for contact with the soldering iron. For prototype work, a hotplate may be used to achieve solder reflow, but a controlled reflow oven process is preferred.

The eight metal leads of the plastic package are also solder coated, so these leads must also be bonded to the circuit-board traces with solder (but not with conducting epoxy). The lead labeled "GND" is connected internally to a grounded metal pad on the prescaler chip. This lead should be soldered to a grounded metal pad on the circuit board.

The backside of the die and the bonding pads on the top side of the die are Au metallized. The 4-mil-thick chips feature 1 mil of backside metal. The bonding pads on the topside of the chip are 4 × 4-mil squares. The die may be attached to a metal carrier using either AuSn eutectic solder or electrically conductive epoxy. Several bonding pads on the prescaler chip (identified as GND) are ground pads that are connected within the chip directly to the backside-grounded metal on the device. With the die attached by AuSn eutectic solder or electrically conductive epoxy directly to a circuit carrier, it is not necessary to run bond wires to these pads. The pads can be used as bonding points for grounding other circuit elements if necessary.

The phase-noise characteristics of a prescaler should be well-understood when the device is used in phase-noise-critical applications. There are three basic processes to keep in mind:

1. Theoretical analysis of a divide-by-N frequency divider shows that the output signal phase noise, expressed on a per-hertz basis and referenced at the same frequency delta from the carrier, is reduced by $20\log_{10}(N)$ dB, compared to the phase noise of the input signal. (Recall that for a ×N frequency multiplier, the phase noise is increased by the same $20\log_{10}(N)$ dB factor.)

2. Due to thermal noise, 1/f noise processes, etc., there will always be some degradation of the output signal phase noise from the prescaler relative to the theoretical value. As seen in Fig. 2, the phase noise that is added by a prescaler to a synthesizer is negligible.

3. Noise on the DC power-supply line can be a significant source of phase-noise degradation. Good device decoupling should always be used, along with proper power-line filtering, shielding, and grounding for the PCB layout.

Examples of plastic-packaged and chip prescalers with connections configured for single-ended input and output operation are shown in **Fig. 3**. For the chip, the three control terminals (input disable, power down, and output-power select) are configured for full output-power operation.

The firm offers evaluation boards, primarily for the plastic-packaged prescalers, although they can also be configured for chip prescalers. The fully assembled boards, which are set up for 50-Ω single-ended input operation include the prescaler, DC blocking capacitors, decoupling capacitors, and surface-mount-architecture (SMA) input and output connectors. A user need only apply an RF input signal and DC power, and RF output signals are available at SMA connectors.

As an example of the type of design possible with these prescalers, a 10.6-GHz phase-locked oscillator (PLO) for a frequency-reference requirement in a Synchronous Optical Network (SONET) OC-192 (10-Gb/s) optical-communications system was constructed **(Fig. 4)**. It is based on two Hittite prescalers: a divide-by-2 unit followed by a divide-by-8 unit. The 10.6-GHz VCO in this circuit is a proprietary design, although other tunable sources, such as a hybrid-circuit varactor-tuned dielectric-resonator-oscillator (DRO) assembly, can be substituted. The HMC403S8G digital phase-frequency detector is a standard Hittite MMIC, developed for high-frequency, low-phase-noise applications. Each of the three MMICs is housed in a surface-mount plastic package. **Figure 5** compared the phase noise of the free-running VCO and the PLL.

Figure 6 shows a block diagram for a microwave frequency synthesizer designed and fabricated using a plastic-packaged model HMC362S8G divide-by-4 MMIC prescaler, along with a commercial fractional-N synthesizer IC from Analog Devices (Norwood, MA). The phase noise of the synthesizer was measured **(Fig. 7)** with a 3048A test set from Agilent Technologies (Santa Rosa, CA). The prescaler delivers an output signal which is sufficiently low in frequency for the synthesizer IC to process. In this circuit, at the point before the active frequency doubler, the circuit is essentially a 13.0-to-13.3-GHz frequency synthesizer with 5-MHz frequency steps. The addition of the ×2 active multiplier turns it into a 26.0-to-26.6-GHz synthesizer with 10-MHz steps. The measured phase noise at 26 GHz is suitable for systems employing 16-state quadrature amplitude modulation (16QAM).

The VCO in this synthesizer is the company's model HMC401QS16G MMIC. The push-push source has an internal signal available that is one-half the output frequency, thereby eliminating a separate divide-by-2 stage. An integrated active doubler delivers the desired 26-GHz frequency range. Active multipliers in this frequency range are also available in hybrid form from a variety of manufacturers. MRF

Continued from page 118
contact member have tapered ends, which produce a spring action and an outward force along a center axis of cylindrical pressure contact.

To achieve low-frequency performance, multiple capacitors are employed in a coaxial structure **(Fig. 6)**. Pairs of parallel-plate microwave capacitors are connected to minimize losses at the lowest frequencies. The outer plates of the two capacitors are connected through a special miniature clip. This configuration enables the total capacitance to be increased and also provides a way to install the capacitor assembly in a coaxial structure. Since the total capacitance increases, the loss at low frequencies is reduced. By using this with a pair of axially resilient coaxial connections on both ends of the center conductors, a solderless connection is created in the coaxial transmission line.[2]

The DC current is injected through Anritsu's tapered coil-type inductor,[3] which provides advantages in broadband resonant-free applications such as bias tees. The diameters of the coil windings are tapered to reduce resonant loss found in typical inductors, which have uniform diameter windings. In a tapered coil, the diameter of each winding is slightly bigger than the next winding, dramatically reducing the resonant losses. Additionally, the coil includes a core made up of a dielectric material containing a colloidal suspension of magnetic particles. This provides low resistive losses and enables the coil to have large low-frequency quality factor (Q). Since magnetic particles will have magnetic permeability, the coil will have an increased inductance at all microwave frequencies. As such, a single coil can be used in a filter, which requires a large low-frequency Q, and as a bias line that requires large resistance at high frequencies for this particular high DC current application. Anritsu Co., 1155 East Collins Blvd., Richardson, TX 75081; (800) ANRITSU, Internet: www.global.anritsu.com/products/components

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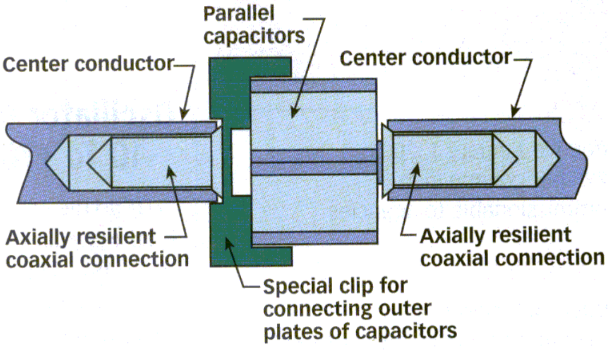

6. To achieve good low-frequency response in the bias tee, multiple parallel-plate capacitors are employed in a coaxial structure.

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Note: The V Connector is a trademark of Anritsu Co.

1. US Patent Nos. 5,576,675 and 6,053,755.
2. US Patent applied for.
3. US Patent applied for.

DESIGN

Continued from page 100
applied to it by a 20-kHz lowpass loop filter. The order of the loop will determine how much attenuation is applied per decade. So, an integer-N inherently attenuates the spur. The integer-N ADF4113 spurious performance in GSM-900 conditions is typically better than −95 dBc.

This is where the fractional-N begins to run into trouble. A 2-MHz PFD will generally have a 200-kHz loop filter. (A 20-kHz filter could be implemented, but this would only result in lock times similar to the integer-N.) There will be a spur at 2 MHz, which will be well-attenuated by the filter, just like an integer-N. But the sigma-delta circuitry will also introduce fractional spurs into the system. As explained earlier, the first fractional spur will usually appear at $F_{PFD}M$, which is the same offset that the first reference spur would appear in an integer-N architecture. Again, its harmonics will also appear. In the example used here, this means a fractional spur will appear at 200 kHz, which is just at the cutoff of the 200-kHz loop filter. Therefore, the filter will not attenuate this spur, and its energy is modulated directly onto the RF output. In practice, this results in fractional spurs at levels that violate acceptable limits for many applications.

What is the best choice of synthesizer for a particular application and why are integer-N synthesizers so widely used? There are several reasons:

- Technology is still battling to overcome the noise and spurious problems inherent in implementing fractional-N sources. The nonlinearity of the charge pump becomes a critical factor in the level of fractional-N spurs.
- Integer-N PLLs are still good enough for many applications and are improving all the time.
- The extra circuitry needed to implement a fractional-N solution increases current consumption. In the wireless portable market, current consumption is a critical specification.
- The "holy-grail" of the PLL designer will appear in the form of a fractional-N synthesizer whose compensation circuitry does not introduce any excess noise or spurs into the system, thereby providing all of the theoretical advantages over integer-N.
- Finally, fractions are more difficult to understand than whole numbers. It was true in school, and is true in the PLL world.

ACKNOWLEDGMENTS

The author would like to acknowledge Mike Curtin, Adrian Fox, Colin Lyden, and the RF group at Analog Devices for their help in researching the article.

FOR FURTHER READING

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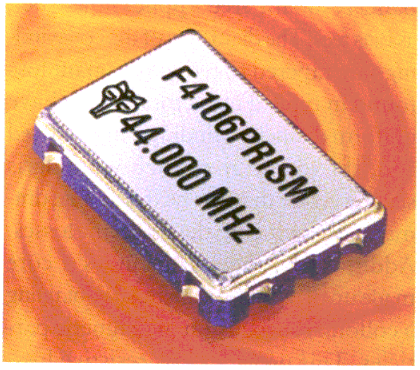

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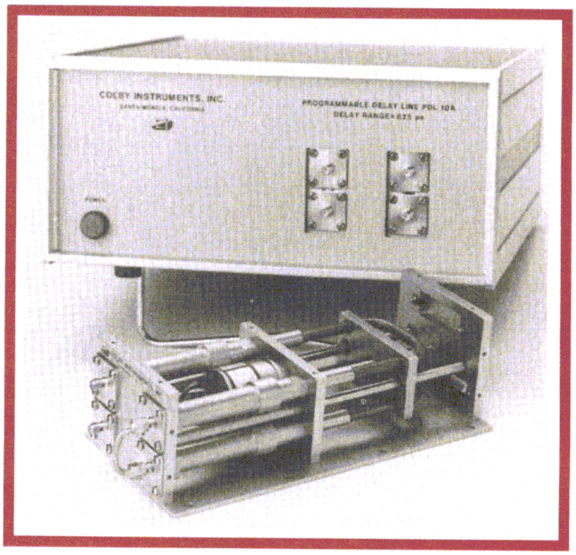


TWELVE YEARS AGO, a Special Report on phase shifters and delay lines introduced the precision of the model PDL 1A programmable delay line from Colby Instruments (Santa Monica, CA), with 625-ps total delay and 1-ps tuning resolution at 1 GHz.

next month

Microwaves & RF October Editorial Preview

Issue Theme: Integrated Circuits

News

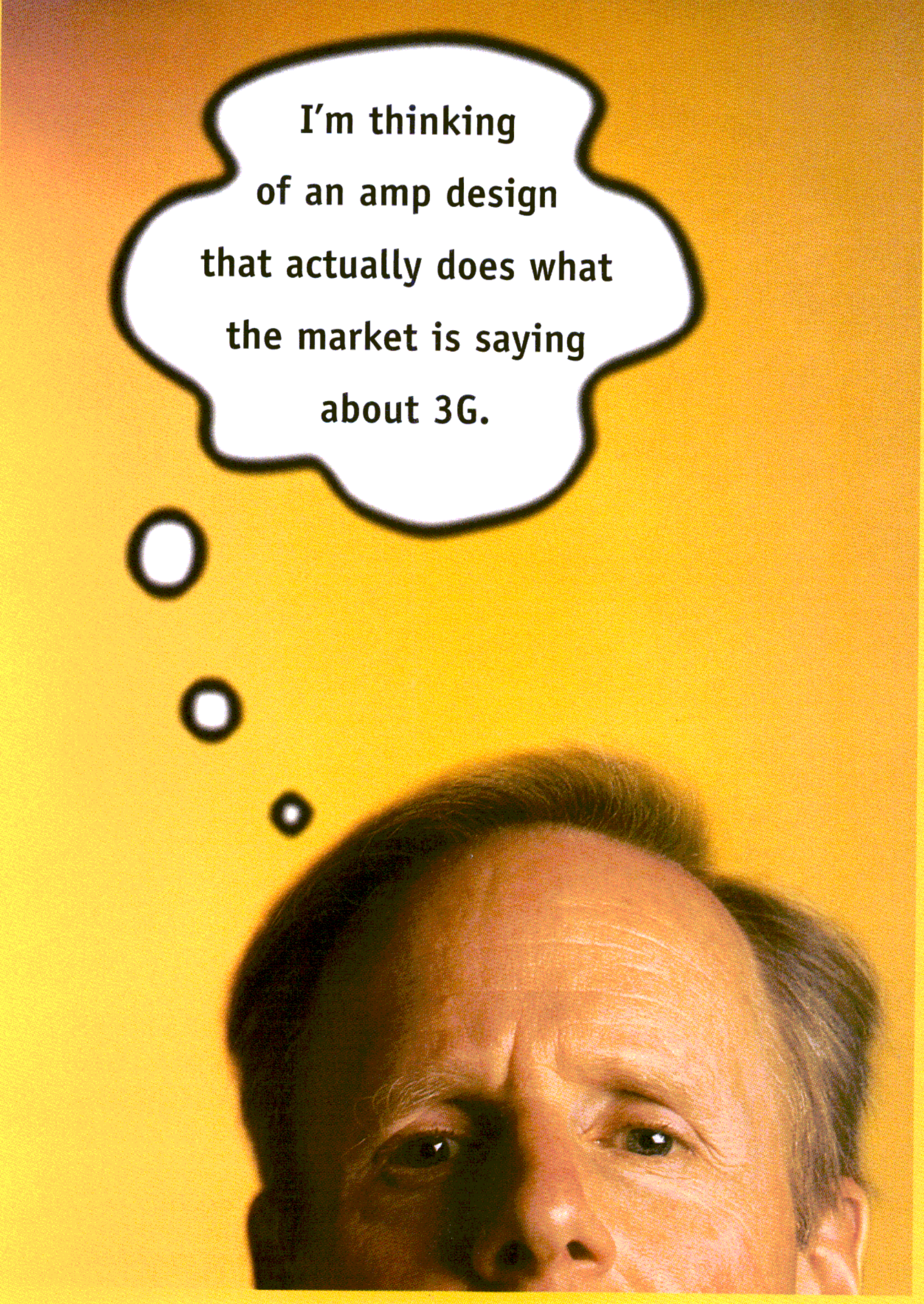
FTIAs, once associated with lower frequencies, are now being designed at higher microwave frequencies in support of high-speed optical-communications systems. What are the limitations of current semiconductor processes for TIAs and how are they being employed in modern optical-communications systems? Who are the key suppliers of high-speed TIAs? Do not miss this special technology update.

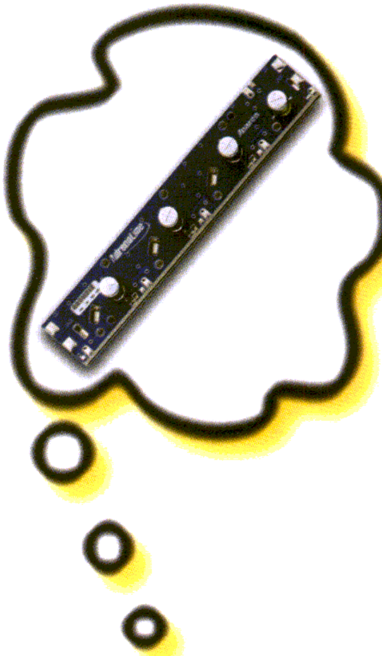
Design Features

October offers a host of design articles on ICs, software modeling, and measurement strategies. For example, an author from RF Micro Devices (Greensboro, NC) explores the definition of efficiency and how to achieve it in IC Tx designs. Additional articles examine the use of harmonic-balance techniques for modeling nonlinear circuits and methods for making triggered measurements with a sampling power meter.

Product Technology

The October Product Technology section will lead off with a novel IC—a polar modulator that provides electronic control of signal phase and amplitude, making it suitable for amplifiers with feedforward linearization and in smart antennas. Other product features will highlight a microwave switch based on MEMS technology, a sensitive Rx based on HTS technology, and a low-power GPS Rx module.





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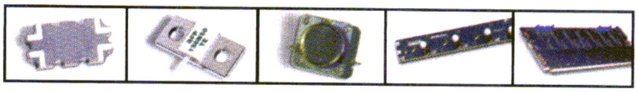

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