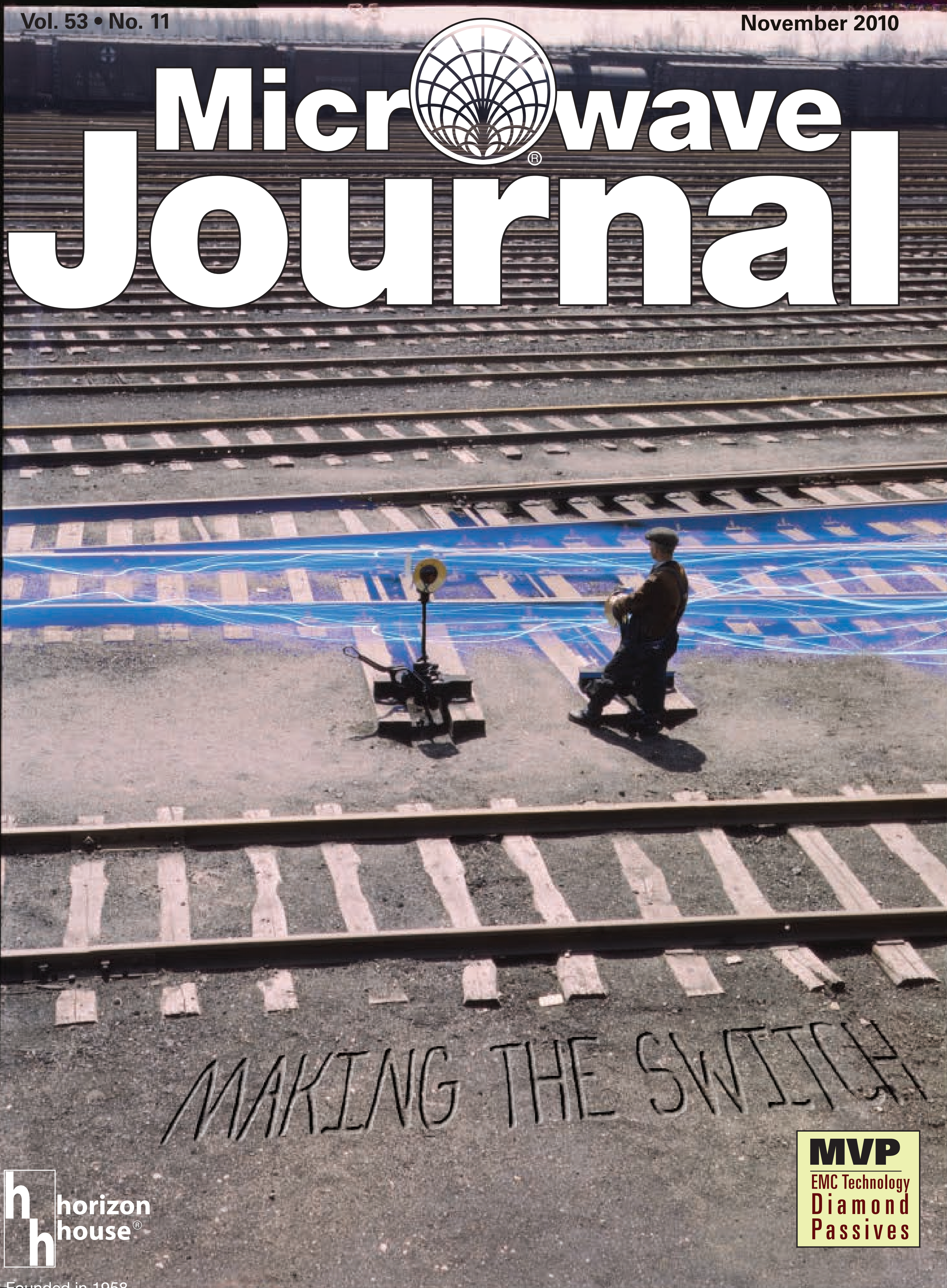
Vol. 53 • No. 11

November 2010

Microwave Journal



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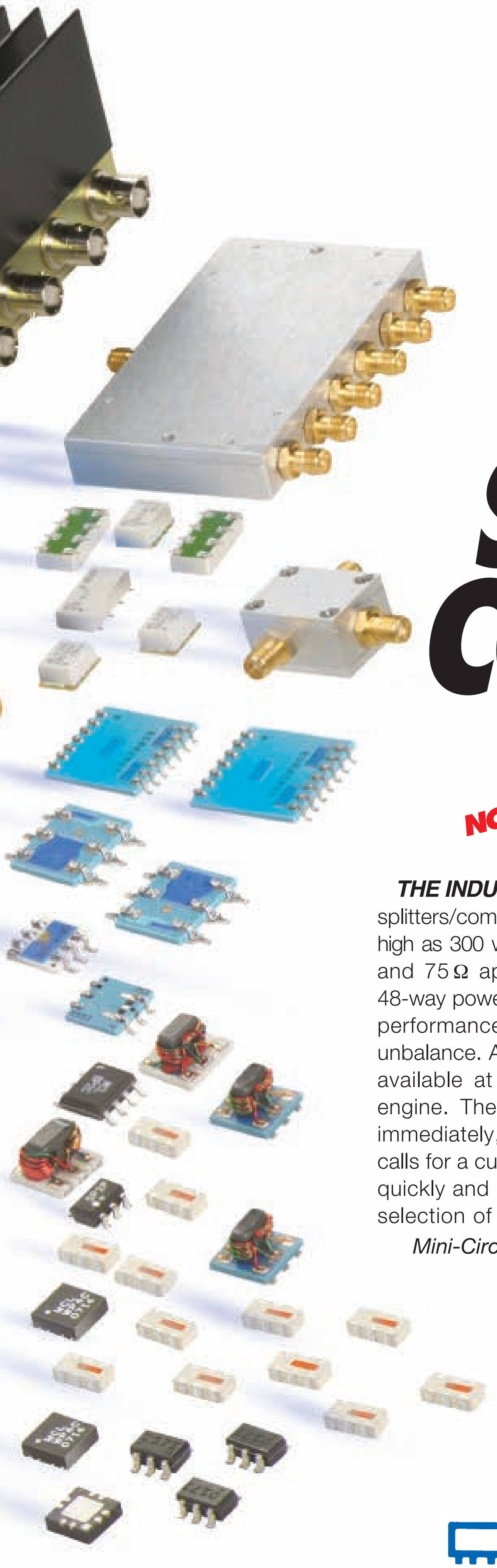


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NOVEMBER 2010 VOL. 53 • NO. 11

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Microwave Journal (USPS 396-250) (ISSN 0192-6225) is published monthly by Horizon House Publications Inc., 685 Canton St., Norwood, MA 02062. Periodicals postage paid at Norwood, MA 02062 and additional mailing offices.

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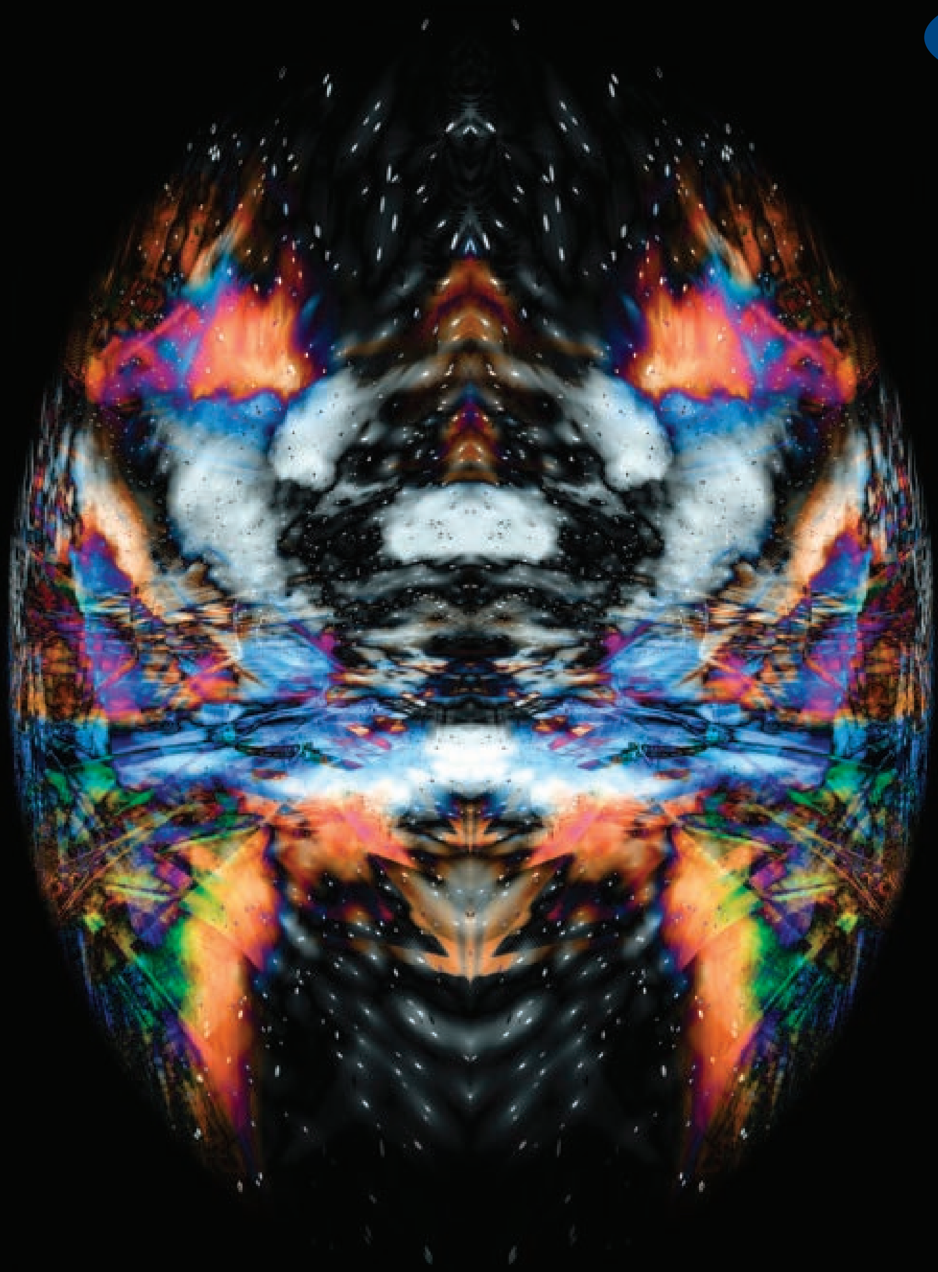

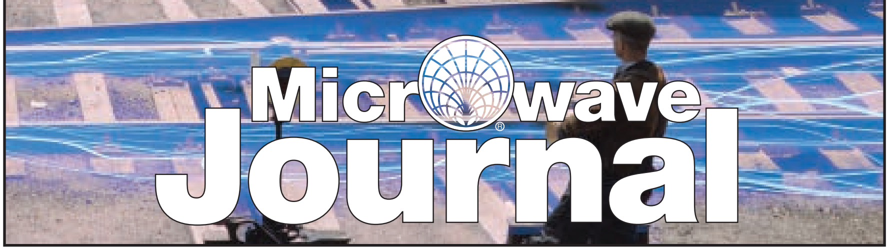

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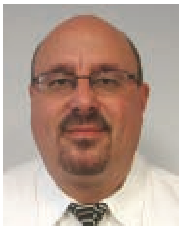

Dale Wildes, Executive Vice President of **KMW Communications Inc.**, discusses his company's origins, global trends among wireless carriers and original equipment manufacturers, and the state of wireless infrastructure technologies.

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Leo Maloratsky, author of *Passive RF and Microwave Integrated Circuits,* has worked in RF/microwave R&D for several decades and is an assistant professor at the Moscow Institute of Radio-Electronics. This month, he contributes engineering tips on transmission line design and applications.

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SUNDAY	MONDAY	TUESDAY	WEDNESDAY	THURSDAY	FRIDAY	SATURDAY
28	29	30	1	2	3 **IMS 2011** Call for Papers Deadline: **December 3, 2010**	4
5	6	7 APMC 2010 **Asia Pacific Microwave Conference** Yokohama, Japan (Dec 7–10) **Webinar: GaAs LNA Design Trade-offs in the Working World** Sponsored by freescale	8 **Webinar: ANSYS 2010**	9 **Webinar: CST STUDIO SUITE 2011 Integrating Simulation Technology** Sponsored by CST	10	11 **ISM 2010**
12 **International Symposium on Microwaves 2010** Bangalore, India	13	14 **MWJ/Besser Webinar: Component Modeling** Sponsored by AWR	15	16	17	18
19	20	21	22	23	24	25
26	27	28	29	30	31 **Next month: Radio Wireless Week** January 16-19, 2011 Phoenix, AZ	1

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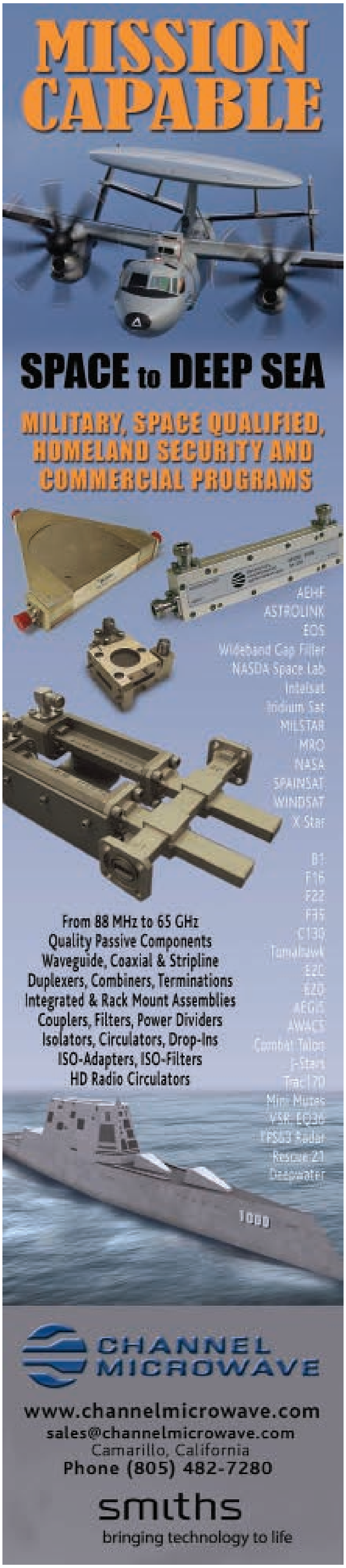

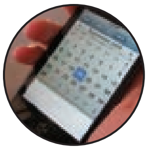

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ARFTG 2011
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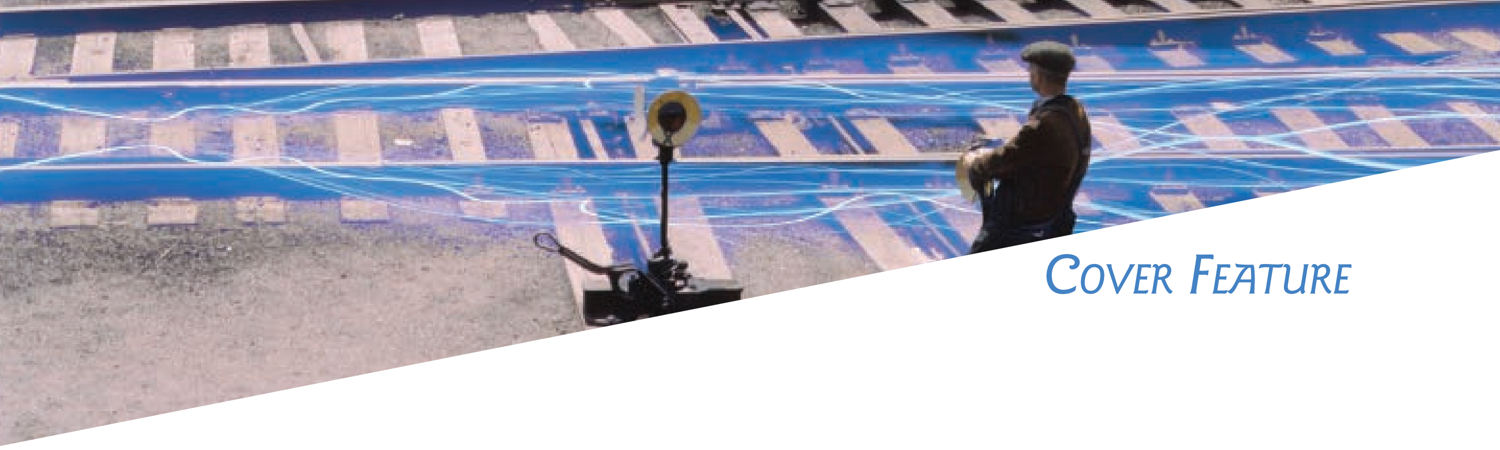

COVER FEATURE

THE STATE OF RF/ MICROWAVE SWITCHES

RF and microwave switches are used extensively in wireless systems for signal routing, finding wide use in switching signals from antennas to the transmit and receive chains. They are one of the highest volume RF devices in use today, as several devices are typically contained in a block diagram. RF and microwave switches fall into the two main categories of electromechanical and solid-state switches. While electromechanical switches have not found wide use in RF and microwave applications since the PIN diode was developed, they are making some new in-roads in certain applications in the form of microelectromechanical systems (MEMS) devices. Solid-state switches are typically more reliable and exhibit longer lifetime than electromechanical switches, plus offer faster switching times. However, solid-state switches typically have higher intrinsic ON resistance and more harmonic distortion than mechanical switches.

Today's CMOS silicon-on-insulator (SOI) and silicon-on-sapphire (SOS) switches are starting to challenge GaAs MMIC switches in many applications as their cut-off frequencies and breakdown voltages improve. Each technology has it advantages and disadvantages, which can be leveraged for various applications for an optimal solution. Up until a few years ago, MEMS switches were considered an emerging technology that had reliability and reproducibility issues. However, recent generations have solved many of those problems making them very competitive in several applications.

OLDIE BUT GOODIE

Starting in the 1950s, PIN diodes were the first widely used solid-state switching technology and are still in wide use today. They excel in very high power and high frequency applications with low insertion loss and better power handling capabilities when compared to most IC FET switches. A PIN diode operates as a variable resistor at RF and microwave frequencies. Its ON resistance varies from less than 1 ohm (ON) to more than 10 kohms (OFF) depending on the bias.[1] One limitation of a PIN diode switch is its lower frequency limit of a few kHz to about a MHz depending on the thickness of the intrinsic or I (intrinsic) region. Therefore, they do not operate all the way down close to DC like most IC switches such as GaAs MMICs or CMOS RFICs. They also require more current to operate compared to IC switches, which means they are typically

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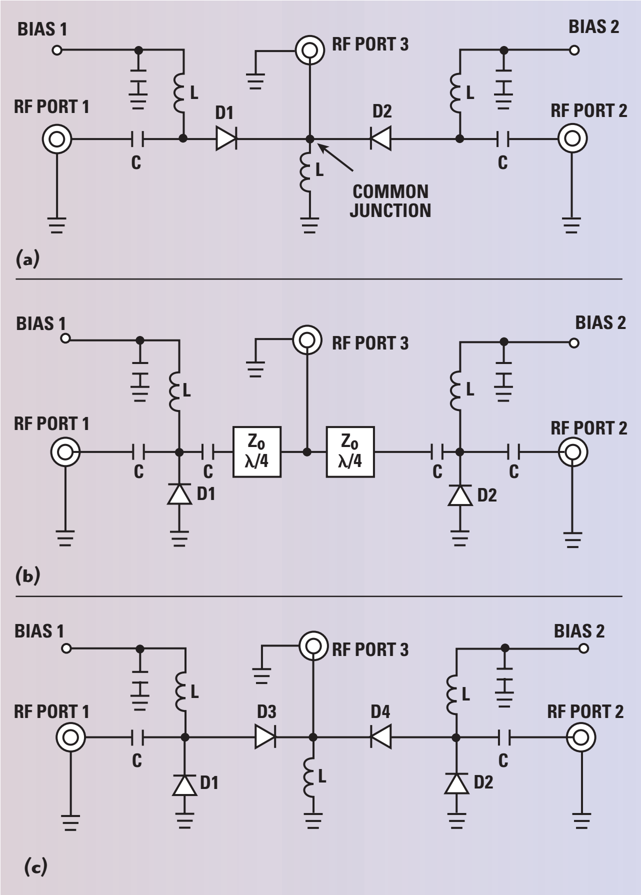


Fig. 1 *Series PIN SPDT switch (a), shunt PIN SPDT switch (b) and series-shunt PIN SPDT switch (c).*[1]

not a good fit for mobile applications. However, they can be more desirable for higher power levels used in military, Satcom or base station applications.

As explained in an Agilent application note,[1] PIN diode switches can consist of a mixture of series and/or shunt diodes depending on the circuit requirements. Series PIN diodes can function within a wide bandwidth limited by the biasing inductors and DC blocking capacitors, while shunt diodes feature high isolation relatively independent of frequency (see ***Figure 1***). Circuit designers often use transmission lines to create series lumped inductance to achieve a low pass filter effect, which enables the switch to work up to the desired frequency. Shunt diode switches have limited bandwidth arising from the use of quarter wavelength transmission lines between the common junction and each shunt diode. A combination of both shunt and series diodes are typically used to achieve optimal insertion loss and isolation performance in a diode switch, but there is a trade off between them. As seen in these various configurations, PIN diodes can require larger circuit areas to realize because of the passive components and multiple diodes needed for the switch design compared to IC switches. The larger footprint is an issue in compact and mobile designs.

Some examples of high power PIN diode products come from Aeroflex/Metelics, a company that offers surface-mount PIN diodes with 100 W CW power handling and 650 W pulsed power handling with insertion loss less than 0.2 dB and isolation of 53 dB. Skyworks offers a series of QFN packaged PIN diodes for high power applications of 50 W CW power handling and 500 W pulsed that operate to 6 GHz with better than 0.45 dB insertion loss and isolation of better than 37 dB. M/A-COM has a special KILOVOLT™ series of PIN diodes (ceramic packaged) that can handle multi-kilo watts of pulsed power for very high power applications in addition to many other offerings. These three companies, along with several others, have been manufacturing PIN diodes for many years, so they are well characterized and have proven reliability in many demanding applications. They are available in many form factors such as chip, beamlead, ceramic packaged and surface-mount packaged in addition to chip scale form factors.

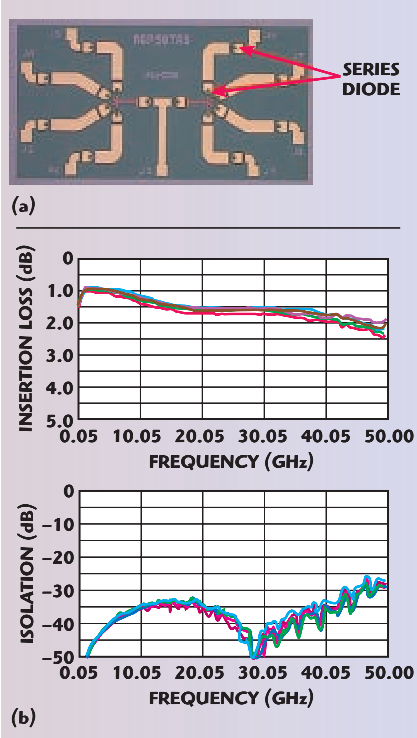
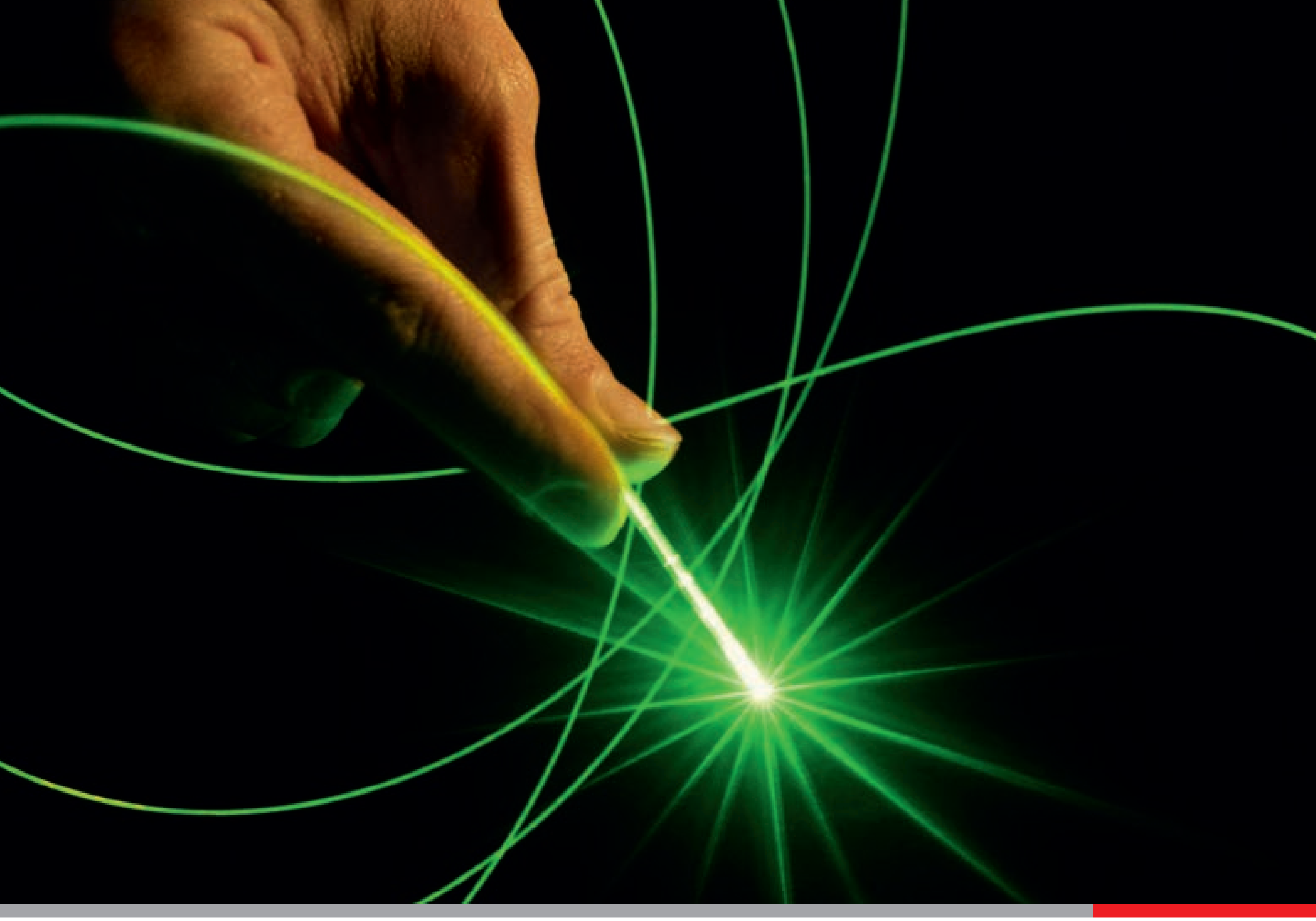


Fig. 2 *An SP8T switch utilizing only series diodes (a) and broadband performance (b).*

PIN diode switches are not just offered as discrete devices; many manufacturers offer integrated diode MMICs. M/A-COM Tech pioneered the diode MMIC in the late 1980s with the Glass Microwave Integrated Circuit (GMIC), which used a glass process to isolate the devices by fusing the GaAs with a glass wafer.[2] This was the predecessor to the Heterolithic Microwave Integrated Circuit (HMIC) process, which is glass and Si fused together. M/A-COM Tech has recently developed multi-octave, high power switches with its AlGaAs PIN diode HMIC devices including multi-throw switches with power handling of 50 W CW (over 100 W pulsed) with less than 2 dB insertion loss and over 30 dB isolation at 40 GHz. This family of switches utilizes a patented AlGaAs heterojunction anode to reduce insertion loss and increase IP3 without compromising isolation; off-state capacitance remains unchanged. This family of AlGaAs/GaAs PIN-based switches probably has the highest frequency, broadest bandwidth response of any of the other technology currently available as these switches are fully functional from 1 MHz to 75 GHz. In addition, switching speeds of less than 500 picoseconds have been measured with this technology. ***Figure 2*** shows a SP8T broadband AlGaAs HMIC PIN diode and its performance.

Another example comes from TriQuint, who offers a vertical PIN process as a foundry service. An example device is a GaAs monolithic PIN diode SP4T switch that operates from DC to 20 GHz. At a bias current of 10 mA per output arm, typical mid-band performance is 0.6 dB



insertion loss with 40 dB isolation in the off-arms. Isolation and insertion loss can be adjusted by varying the output arm bias current of the switch.

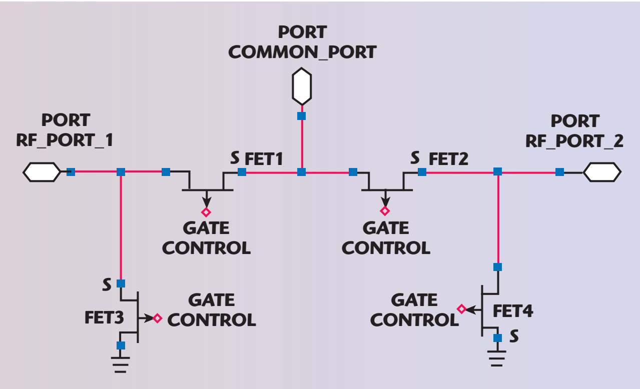
THE CURRENT WORKHORSE

GaAs field-effect transistor (FET)-based switches have been the mainstay of RF/microwave switches since the 1980s when MMIC circuits became widely available at relatively low prices. Driven by DARPA funding for defense applications (MIMIC program) and a high demand for commercial wireless devices, GaAs MMIC reproducibility improved dramatically during this period and device costs were greatly reduced.[3] FET switches are very stable and repeatable due to good control of the drain-to-source resistance. FET switches are voltage-controlled resistors so they provide low power operation, small size and relative design simplicity compared to PIN diodes. They are broadband (DC to 20 GHz devices are widely available) and have relatively high linearity.

Initially, GaAs MMIC MESFETs were widely used in the 1980s and '90s, but these have given way to PHEMT MMIC devices, which have better ON resistance (Ron) characteristics and are now the most widely used GaAs MMIC switching devices. While MESFET devices were able to reach switching speeds down to tens of pico seconds, PHEMT devices suffer from gate lag as electrons can be trapped on the surface. PHEMTs typically have switching times in the hundreds of micro seconds as they can switch in tens of nano seconds (10 to 90 percent), but have gate lag times of several hundred microseconds (90 to 98 percent).

However, new developments such as M/A-COM's nanosecond designs achieve about ten nano second switching times including gate lag. M/A-COM made a number of changes to the PHEMT process and device structure to overcome this problem.[4] The number of surface states and interface traps were reduced at the ungated GaAs surface using cleaning techniques and the deposition of a special passivating dielectric. The formation of the Schottky diode gate was modified to both reduce gate resistance with no additional gate capacitance in order to minimize the RC charging time associated with device turn-on and turn-off. And a proprietary III-V layer was added to the PHEMT structure to further reduce the channel resistance and enable enhanced movement of charge through the device especially from the ungated recess region. This process optimization for low gate lag not only resulted in an improvement in the 90 to 98 percent switch settling time, but also exhibited reduction in the 10 to 90 percent switching speed. While CMOS Si-based switches do not suffer from gate lag, they typically switch in the range of micro seconds. This is because CMOS SOI devices are typically designed for low-frequency operation. Similarly, SOS switches can achieve nanosecond switching speeds without gate lag; however, most devices have been designed to optimize the tradeoff between speed and low frequency operation. One of the drawbacks of MEMS switches is switching speed as they typically exhibit speeds of tens to hundreds of micro seconds for electrostatically operated devices.

Low channel resistance allows GaAs MMIC switches to operate at low frequencies (very near DC) and reverse biasing completely depletes the channel in the OFF state providing excellent isolation at low frequencies.[1] However, to operate down near DC, GaAs switches do need a charge pump circuit because the traditional capacitors prevent low frequency operation. The isolation degrades at higher frequencies due the effect of the drain-to-source capacitance. ***Fig-***



▲ *Fig. 3 Simplified SPDT switch using FETs as switching devices.[1]*

ure 3 shows a simplified schematic diagram of a SPDT FET switch. The biasing path is not connected to the RF path simplifying the DC biasing path and eliminating the expensive RF choke. The chokes are used to reduce the insertion loss that results from the biasing path being connected to the RF port in PIN diode switches. The ON resistance of a GaAs FET is still typically higher than a PIN diode, so the insertion loss performance of FET switches is not as good as PIN switches. FET switches are voltage controlled so they consume far less current than current controlled PIN switches.

While GaAs MMIC switches were originally only available as depletion mode devices, they required a negative control voltage for operation that was not desirable compared to CMOS switches. An alternative to using a negative control voltage is to elevate or float the DC voltage at the source of the FET to +5 V and use a 0 to +5 V control voltage. Floating the DC voltage requires blocking capacitors that complicates the design and requires more circuitry. However, there are now a number of suppliers offering enhancement mode PHEMT (E-PHEMT) devices that do not require a negative gate voltage to operate (these devices are typically only offered in PA MMICs). They are normally OFF and use a positive voltage to turn the FETs on. This also allows integration of limited logic on the same chip that has always been an advantage with Si-based FET switches, but it is at the cost of current consumption. Many companies and foundries also offer E/D-PHEMT processes that can incorporate both FET modes on a MMIC so that each device type within the circuit can use the process that best fits its needs for performance. Companies such as TriQuint, RFMD, Skyworks, Hittite and M/A-COM Tech, along with others, use this technology in their switches for appropriate applications where the increased complication in processing is worth the added benefits. Therefore, the disadvantages for negative voltage operation and logic integration for GaAs MMICs has been decreased over the past few years, although Si CMOS still offers better integration opportunities with logic and memory circuits.

The GaAs MMIC switch market is still very large and they are used in many applications from commercial to military. Many circuit designs and switch types are available and optimized for almost any application. The design demands for compact, low current multi-throw handset devices are far different from high power base station or military radar applications. Most of the widely known component manufacturers such as Skyworks, RFMD, TriQuint, Hittite, CEL/NEC, Mini-Circuits and M/A-COM Tech offer a wide variety of devices depending on the application. GaAs MMIC devices are still progressing as they shrink die sizes, develop chip scale packaging and optimize the FET design, but the technology is relatively mature so the improvements are not revolutionary at this point. Over the last 20 years or so, GaAs PHEMT switches have offered the best overall performance for most high frequency (over a few GHz) and broadband applications that require low to medium power levels, but this is changing.

GAINING GROUND

Standard Si CMOS-based FET switches have previously not proven to be good RF switches as they suffered from significant insertion loss and low isolation because the substrate is not insulating and breakdown voltages are low. One way to overcome the low breakdown voltage is to stack the FETs, but it is difficult to accomplish spreading the voltage evenly across the FETs so this has not worked very well in standard CMOS. However, Si-on-sapphire (SOS) and more recently Si-on-insulator (SOI) FET switches have been gaining market share in many applications as their insulating substrate quality, cut-off frequencies and breakdown voltages have improved. They are competing with GaAs switches in some high volume applications such as handset switches and even demanding military applications.

These Si technologies accomplish FET stacking with even voltage distribution across the FETs, low insertion loss, high isolation and better linearity than standard CMOS. An example of higher voltage operation is Peregrine Semiconductor SOS switches that have recently achieved 50 W

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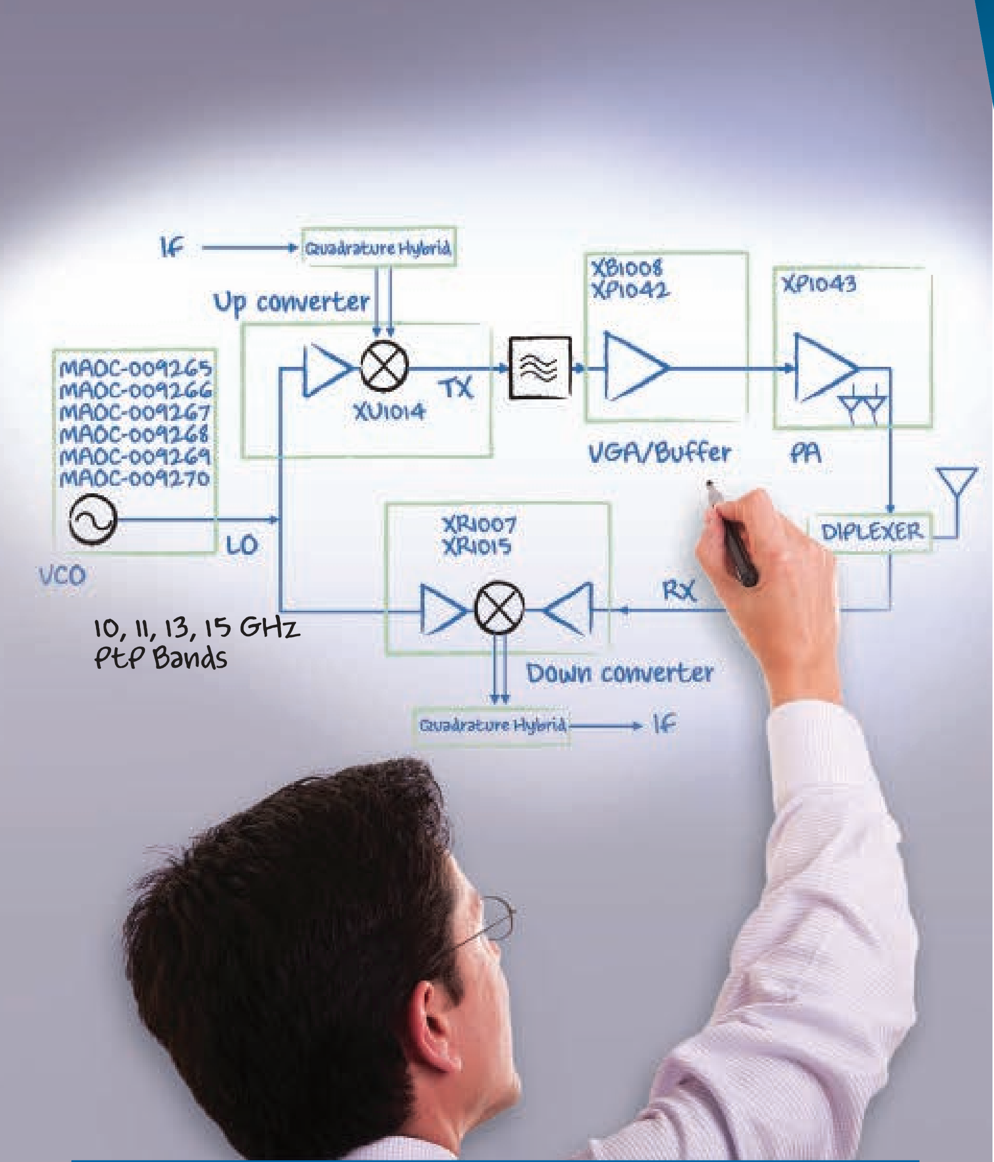

Part Number	Frequency (GHz)	Gain (dB)	P1dB (dBm)	OIP3 (dBm)	Pout (dBm)	PN at 100 kHz (dBc/Hz)	Current (mA)	Package (mm)
XP1043-QH	12.0-16.0	21.5	30.0	41.0			700	4x4
XP1042-QT	12.0-16.0	21.0	25.0	38.0			500	3x3
XB1008-QT	10.0-21.0	17.0	19.0	32.0			100	3x3
XU1014-QH	8.0-18.0	-10.0	2.0	12.0			80	4x4
XR1007-QD	10.0-18.0	13.5		4.0 (I/P)			150	7x7
XR1015-QH	10.0-16.0	12.0		2.0 (I/P)			170	4x4
MAOC-009265	9.0-10.3				6.0	-110	175	5x5
MAOC-009266	10.2-11.3				9.0	-114	185	5x5
MAOC-009267	11.2-12.6				3.5	-110	165	5x5
MAOC-009268	12.7-14.2				7.0	-105	175	5x5
MAOC-009269	11.4-12.8				3.0	-110	165	5x5
MAOC-009270	12.2-13.8				6.5	-105	155	5x5

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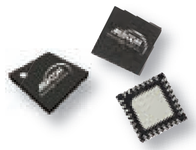

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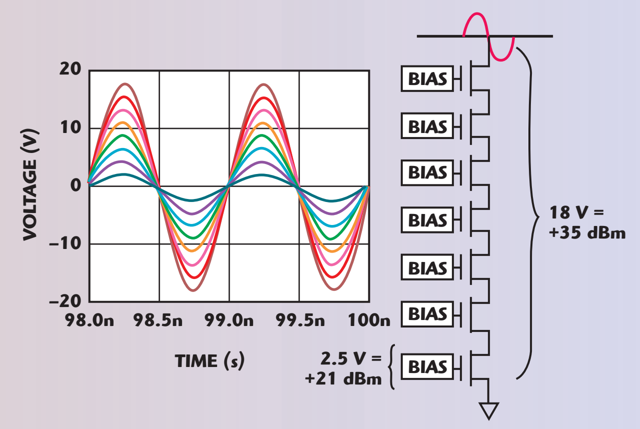
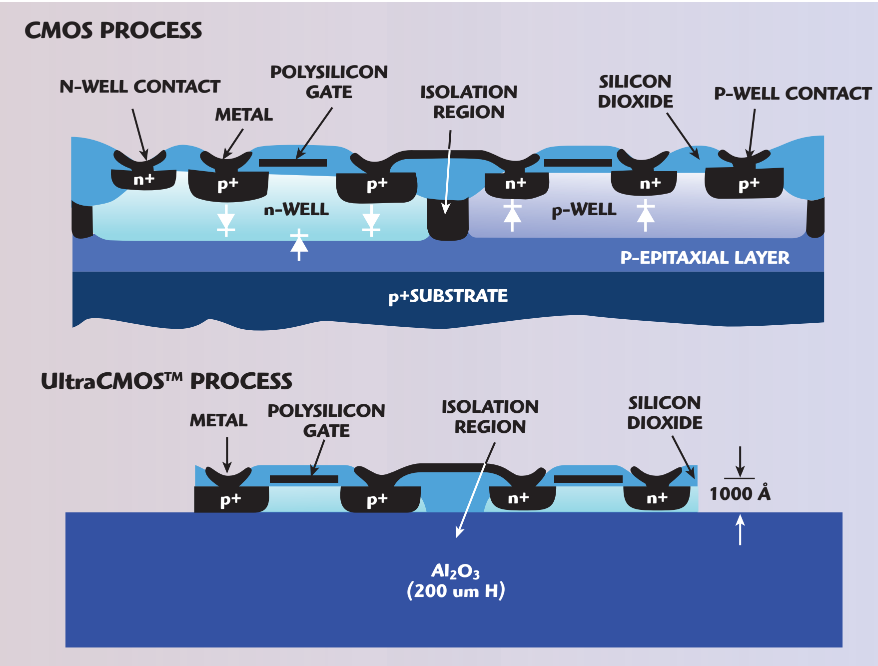
CW power handling and greater than 80 dBm IIP3 on switch designs that are much higher than GaAs MMIC switches and similar to PIN diode MMICs (see ***Figure 4***). This is allowing them to penetrate even the high power territory where PIN diodes have dominated. Peregrine maintains that SOS provides the highest linearity due to its fully insulating substrate and GaAs linearity has limitations due to the diode junction formed between the gates in the channel.

Peregrine Semiconductor and IBM recently teamed up to develop and manufacture future generations of Peregrine's patented UltraCMOS™ silicon-on-sapphire (SOS) process technology, which is unique for its thin insulating layer. It provides the needed isolation, but is thin enough so that it minimizes the negative effects of a thicker Si layer that does not provide ideal high resistivity (see ***Figure 5***). When fully qualified, the next-generation UltraCMOS RF ICs will be manufactured by IBM for Peregrine in the jointly-developed 180-nanometer RF CMOS process at IBM's 200 mm semiconductor manufacturing facility in Burlington, VT. This development marks the first commercial use of 200 mm (8-inch) wafer processing for a silicon-on-sapphire process. An example of a high performance switch recently developed by Peregrine is a monolithic symmetric SP8T switch (manufactured on its STeP5 process) that covers from near DC to 4 GHz with IIP3 of +70 dBm, IIP2 of +130 dBm and insertion loss of 0.35 dB (900 MHz). The switch handles +35 dBm operating input power (across the range) with high ESD tolerance of up to 4 kV (HBM).

While GaAs MMIC switches offer good linearity and isolation with low ON resistance and low OFF capaci-



▲ *Fig. 4 CMOS FET stacking (courtesy of Peregrine Semiconductor).*



▲ *Fig. 5 Traditional CMOS construction (a) and UltraCMOS™ construction (b).*

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tance (Coff), they do have some disadvantages. GaAs technology is relatively mature and while it still is improving, most major advances have probably been achieved. As a representative from Peregrine said, "There are not many more dials to turn to improve GaAs device performance." SOS and SOI CMOS-based devices are improving quickly and are now closing in on a lower Ron·Coff product, a good figure of merit for switches, allowing the design of switches with lower insertion loss and higher isolation. Peregrine believes that SOS could achieve a product of less than 200 fs as they progress to 0.18 micron technology and even lower with 0.13 micron technology. Today some high performance GaAs devices are already at this level.

Up until recently, SOI resistivity was not as ideal as GaAs or SOS, so devices made with SOI technology exhibited higher levels of harmonic and intermodulation distortion. However, recent advancements in SOI CMOS technology have been able to reduce these effects to make them competitive with SOS.[5] RFMD has demonstrated a SP9T SOI switch with similar performance to current PHEMT switches with a Ron·Coff product of 250 fs, which is close to a high quality PHEMT value of 224 fs. This compares favorably to current SOS products of 400 fs with 0.25 micron technology, but with 0.18 micron technology SOS is expected to be below 200 fs.

RFMD has three new SOI switches that are being released. This lineup has been designed into two major handsets, so it seems to be gaining momentum. Earlier this year, Skyworks introduced a symmetrical SP4T SOI switch. The device is designed for 3GPP bands from 0.70 to 2.7 GHz with typical insertion loss as low as 0.6 dB and isolation as high as 30 dB and harmonic performance less than 75 dBc at 0.9 GHz. At the same time, the company also introduced a W-CDMA DP4T SOI switch with a decoder. Today's best GaAs PHEMT switches, 0.18 micron SOS and 0.18 SOI appear to be nearly equal in this figure of merit measurement. ***Table 1*** compares the Ron·Coff figure of merit for various technologies. It should be noted that this figure of merit is effective for comparing raw device performance, but is not the only metric important for a complete switch design. Therefore, processes with lower Ron·Coff values can outperform ones with lower values when other tradeoffs are made in a complete design.

GaAs MMIC devices can have higher contact resistance than these Si technologies, increasing losses, and cannot integrate logic circuits as well as Si-based technologies. Driving GaAs switches also frequently requires extra interface components, and GaAs has limited capability to integrate other functions such as logic control and memory. While CMOS switches have greater than 2000 V (HBM) ESD tolerance, which is relatively robust, most GaAs MMIC switches are only around 200 V, making them susceptible to ESD damage and typically requiring special handling procedures. The same is true with most RF MEMS switches.

Cost is a major advantage of any type of Si wafer processing for large volume applications compared to GaAs because Si has lower material

TABLE I

FIGURE OF MERIT FOR VARIOUS SWITCH TECHNOLOGIES

Process	Device	Ron (ohm-mm)	Coff (fF/mm)	Ron·Coff (fs)
0.18 thick film SOI	5 V NFET Lg = 0.6 μm 13 nm gate oxide	1.9	255	485
0.18 thin film SOI	2.5 V NFET Lg = 0.32 μm 5.2 nm gate oxide	0.8	310	250
0.5 μm SOS	NFET 10 nm gate oxide	2.8	270	756
0.35 μm SOS	NFET 10 nm gate oxide	1.1	240	264
GaAs PHEMT	R-JPHEMT	1.4	160	224

costs and larger wafers to reduce the cost per unit area. While in the past SOS substrates were very expensive, this is changing as the LED market is fueling demand for lower cost substrates and driving high volumes. According to Peregrine Semiconductor, government funding for the LED market along with the demand for low energy lighting could make sapphire substrates the highest volume electronic devices in the near future (see ***Figure 6***). LEDs have been running on 150 mm substrates, putting them on the same wafer sizes as high volume GaAs.

However, neither SOS nor GaAs wafer processing can probably come close to the low cost of SOI switches, which use standard Si processing and larger wafer sizes, although Peregrine maintains that its SOS process uses fewer masking steps and standard processes that can match SOI costs. As SOI performance improves to match GaAs and SOS switches, they will probably have a cost advantage for high volume applications. One way that GaAs devices compete on cost is their setup costs (lower cost mask sets) are typically much less for wafer runs than Si, so for lower volumes they can gain a cost advantage especially for IDMs that produce their own devices.[3]

While SOI technology has not been competitive in the past, it seems poised to compete in the lower frequency (under 3 GHz), high volume market such as handset and perhaps even WLAN markets. This is mostly enabled by the recent availability of low cost, high resistivity Si substrates (1 kohm-cm), which were not available several years ago. As one representative from RFMD put it, "While SOI was thought to have poor linearity, we are finding that through careful switch branch layout, charge pump optimization and an excellent collaboration with our foundry partners, we are now meeting or exceeding the best of SOS reported performance, and SOI substrate costs are a fraction of sapphire at 200 mm, not counting the fact that SOI uses standard technologies and libraries." Peregrine conversely states that its SOS technology uses standard CMOS processes and libraries. The company also maintains that its Step5 process leads the market in all performance parameters and provides better design flexibility.

STARTING TO COMPETE

RF MEMS capacitive switches were first developed and used in the early 1990s and typically use an electrostatic means to actuate the switch. They offer very low loss and high linearity compared to FET switches, but their switching speed is typically much lower. There are basically two types, the ohmic contact and capacitive contact. With ohmic switches, two metal electrodes are brought together to

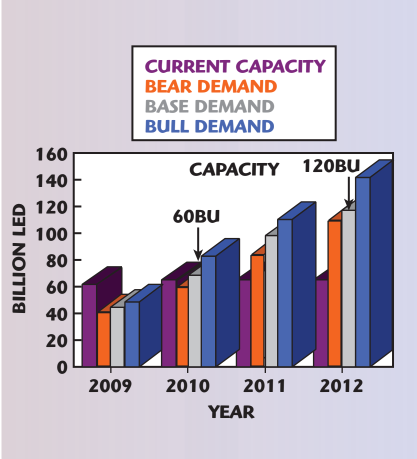

Fig. 6 Projected demand for sapphire substrates (source: Canaccord Adams estimates).

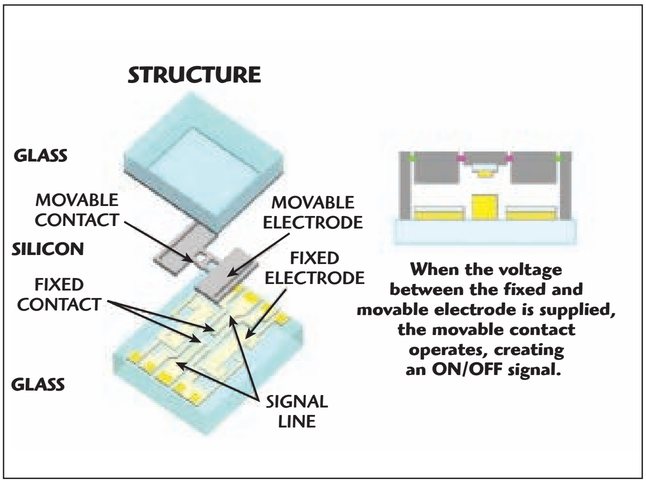

Fig. 7 Example capacitive MEMS switch construction (courtesy of Omron).

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RFSP5522*	5 to 1200	14	22	18	1.0	1.6	0.6	5.0	S18
RFSP5722	5 to 1200	12	28	22	0.8	1.5	0.5	5.0	S18

* 50 ohms

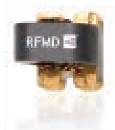
COUPLERS

Part Number	*Freq Range (MHz)*	*Nominal Coupling (dB)*	*Coupling Flatness (dB)*	*Mainline Loss Typ (dB)*	*Mainline Loss Max (dB)*	*Directivity Typ (dB)*	*Directivity Min (dB)*	*Return Loss Min (dB)*	*Package*
RFCP5742	5 to 1200	10 ± 0.5	± 0.5	1.5	2.0	14	10	11	S18
RFCP5743	5 to 1200	10 ± 0.5	± 0.5	1.5	2.0	14	8	14	S20
RFCP5762	5 to 1200	16 ± 0.5	± 0.5	0.6	1.2	20	10	14	S18
RFCP5763	5 to 1200	16 ± 0.5	± 0.5	0.8	1.2	20	8	14	S20

TRANSFORMERS

Part Number	*Freq Range (MHz)*	*Impedance Ratio*	*Insertion Loss 3 dB Bandwidth (MHz)*	*Insertion Loss 2 dB Bandwidth (MHz)*	*Insertion Loss 1 dB Bandwidth (MHz)*	*Package*
RFXF2513*	0.3 to 200	1:4	—	—	3 to 200	S20
RFXF2533*	2 to 350	1:2	—	—	2 to 350	S20
RFXF2553*	1 to 350	1:4	—	—	1 to 350	S20
RFXF2713	5 to 200	1:1	—	—	5 to 200	S20
RFXF3553*	2 to 1000	1:4	—	2 to 1000	2 to 400	S20
RFXF5702	5 to 1200	1:1	—	5 to 1200	5 to 750	S18
RFXF5703	5 to 1200	1:1	—	5 to 1200	10 to 870	S20
RFXF5704	5 to 1200	1:1	—	5 to 1200	5 to 800	S21
RFXF5712	5 to 1200	1:1	—	5 to 1200	5 to 1000	S18
RFXF5753	5 to 1200	1:4	5 to 1200	5 to 870	—	S20
RFXF5792	5 to 1200	1:1	—	5 to 1200	5 to 1000	S18
RFXF5793	5 to 1200	1:1	—	—	5 to 1200	S20
RFXF5794	5 to 1200	1:1	—	—	5 to 1200	S21
RFXF6553*	10 to 1900	1:4	10 to 19000	10 to 1000	10 to 500	S20
RFXF8553*	500 to 2500	1:4	500 to 2500	500 to 1500	500 to 1000	S20
RFXF9503*	3 to 3000	1:1	3 to 2700	3 to 2400	3 to 1800	S20
RFXF9504*	5 to 3000	1:1	5 to 3000	5 to 2700	5 to 1200	S21

* 50 ohms



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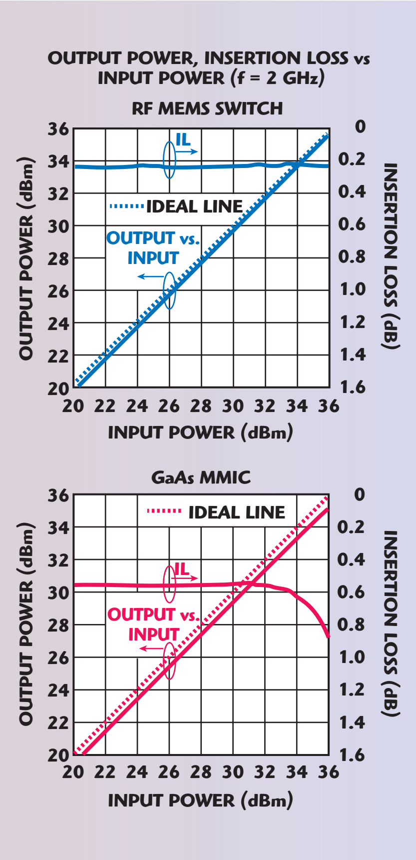
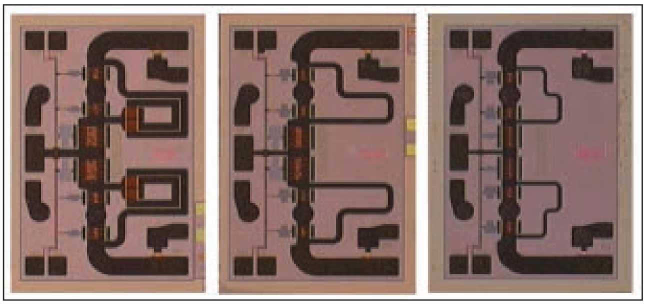


▲ *Fig. 8 RF MEMS switch has no compression point up to +36 dBm (4 W) RF power.*

create a low resistance contact, while in capacitive switches, a metal membrane is pulled down onto a dielectric layer to form a capacitive contact. ***Figure 7*** shows the construction used by Omron for its MEMS switches. The electrode is a special metal composition that flexes down when voltage is applied to turn the switch on and returns back to its original position without the applied voltage. The use of capacitive coupling has reduced issues associated with older generation MEMS switches of dry contact, metal to metal ohmic switching. Issues with sticking contacts, wear, etc., have been mitigated using this newer technology as suppliers have optimized the metallic materials and design.

RF MEMS switches can have no compression point until +36 dBm, as shown in the comparison to GaAs switches in ***Figure 8***. Omron has designed a SPDT switch that operates at 34 V with typical insertion loss of 1 dB, isolation of 30 dB and return loss of 10 dB at 10 GHz. Radant manufactures some high isolation, low loss MEMS switches such as a SP6T DC to 20 GHz device with 22 dB isolation and less than 0.8 dB loss at 18 GHz and near zero harmonic distortion. They also have a very high isolation DC-12 GHz MEMS switch with better than 70 dB isolation and less than 0.3 dB insertion loss at 2 GHz.



▲ *Fig. 9 Die photographs of TriQuint's 6, 12 and 18 GHz SPDT switch MMICs.*

Over the past decade, processing improvements, materials refinements and design changes have enabled designs with less than 0.1 dB loss through 40 GHz, low power consumption of tens of nanojoules per cycle and high linearity of greater than 66 dB, according to Memtronics. Reliability is on the order of 100 million cycles, minimum. In fact, Radant Technology reports that its devices have been independently tested by the Department of Defense (DoD) laboratories under a DARPA program to 1.5 trillion cycles, which was conducted in 30 months of continuous testing. This has allowed them to compete in several applications such as test & measurement and switching arrays for antennas. The advantage of the mechanical switch is that when it is off, it is physically isolated so there is little leakage. Leakage current is about 100 fA at 100 VDC. MEMS offer lower off-state capacitance and better off-state RF isolation than either FETs or PIN diodes. Like GaAs FETs, they have low ESD tolerance of around 100 V HBM so they require special handling.

iSuppli recently reported that it anticipates RF MEMS revenue to rise to $8.1 M this year, $27.9 M in 2011 and then $223.2 M in 2014. Much of this is projected to be from cell phone front-end adoption of tuning using RF MEMS switches and varactors to help boost the performance of smart phones. iSuppli states that WiSpry and TDK-Epcos are offering RF MEMS for high volume cell phone applications, while Analog Devices, Radant Technologies and XCOM Wireless (in cooperation with Teledyne Technologies) as well as Omron are targeting high end applications for testing and instrumentation such as ATC and RF test. Also noted at EuMW 2010 this year was DelfMEMS, who is also manufacturing and supplying MEMS switches.

An example of the new MEMS tuning technology is the TDK-EPC and WiSpry tunable modules that quickly switch in and out various values of capacitance to dynamically tune handset antenna to maximize efficiency. Start-ups Radant and MEMtronics are also focusing on defense applications. Outside of cell phone and instrumentation, wireless infrastructure switches could be replaced by cheaper, higher performance RF MEMS devices. Another opportunity is in defense applications for radio systems and phased-array antennas.

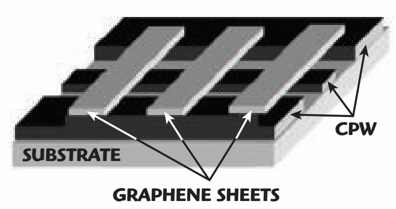
FUTURE TECHNOLOGIES

While most of the recent advances in switch technology have concentrated on the lower power applications driven mostly by handsets, the higher power applications are still dominated by PIN diode technology. GaN has been developed mostly for high power amplifier applications, but it also shows great promise as a future switch technology. A couple of GaN suppliers have started to release switch products. TriQuint has developed three broadband GaN on SiC MMIC switches to cover frequency ranges of DC-6, DC-12 and DC-18 GHz, as shown in ***Figure 9***. These devices have maximum insertion loss of 0.7, 1.0 and 1.5 dB, and demonstrate 40, 15 and 10 W RF power handling, respectively, for 6, 12 and 18 GHz designs.[6] Cree has advanced datasheet information available on a 25 W, 0.1 to 3 GHz SPDT GaN MMIC switch. It features less than 0.7 dB insertion loss, 15 ns switching speed, over 30 dB isolation and over

TABLE II

COMPARISON OF SWITCH DEVICE PARAMETERS

Switch Technology	Monolithic PIN diode	GaAs MMIC	CMOS SOI	CMOS SOS	RF MEMS	GaN MMIC
Insertion Loss (dB)	0.3-1.5	0.3-2.5	0.3-2.5	0.2-1.5	0.1-1.0	0.1-1.5
Isolation (dB)	>=30	>=25	>=30	>=30	>=30	>=30
Power Handling (W)	<=50	<=10	<=50	<=50	<=10	<=100
Power Consumption	High	Low	Low	Low	Low	Low
Switching Speed	ns to μsec	ns to μsec	μsec	ns to μsec	μsec	ns
Ron·Coff (fs)	100-200	224-400	250 (.18 μ)	264 (.35 μ)	100-200	400
Cost	High	Low	Very Low	Very Low	Medium	High
ESD Tolerance	Medium	Low	High	High	Low	High



▲ *Fig. 10 Top view of the graphene NEMS switch configuration.*[7]

60 dBm TOI. With the need for high power switches with lower current consumption, GaN switches should eventually find their way into several applications, especially satellite and military designs.

There also has been some work done with microwave nano electro-mechanical systems (NEMS) switches that potentially could overcome the drawbacks of the current MEMS devices. Work has been done that shows graphene flakes that can operate as a switch up to 60 GHz with switching times of less than a nano second.[7] This could enable all the benefits of MEMS switches while obtaining fast switching times comparable to the fastest solid-state switches. The simplified construction is shown in ***Figure 10***. The device is a coplanar waveguide and an array of metallic graphene sheets suspended over it. The waveguide is made from gold strips deposited on a 500 micron thick semi-insulating Si substrate. The graphene flakes are suspended over the waveguide due to van der Waals forces, but could be attached via metallic contacts.

THE STATE OF SWITCHES

Table 2 shows a summary of the key performance metrics for the various switch technologies covered. The "state" of RF and microwave switch technology today shows that the PIN diode is still very viable in high power, high frequency applications, but the most widely used technology is the GaAs MMIC, which still offers the best performance for most high frequency (over a few GHz) and broadband applications that require low to medium power levels. However, the market is changing as SOS and SOI CMOS switches are making significant in-roads in some high volume applications. Their performance is matching that of GaAs MMICs at frequencies up to a few GHz and they offer cost and integration advantages. SOI and SOS switches are also proving they can be viable in medium to high power applications as breakdown voltages have improved with FET stacking. In the future, GaN MMIC switches show great promise to take a foothold in higher power applications, probably replacing PIN diodes in some of these areas. MEMS switches are showing promise in the test & measurement, phased array and tunable module market, which promises to be significant in the large handset market. ■

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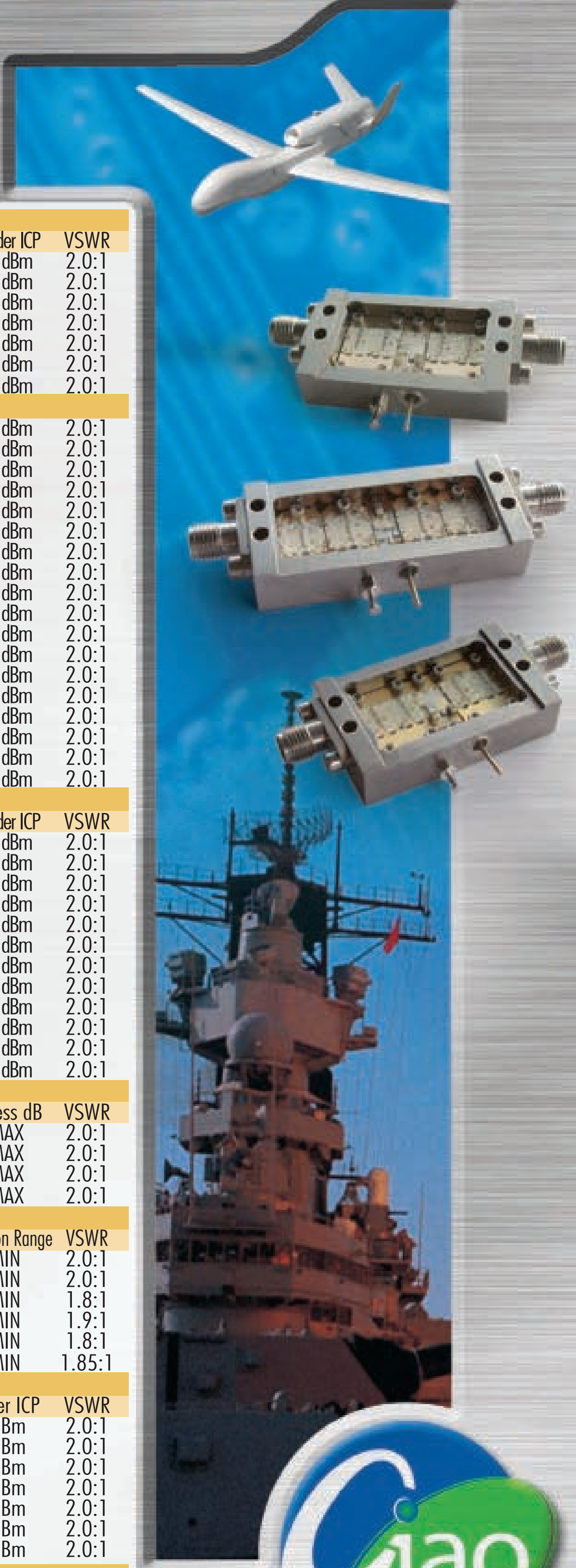

OCTAVE BAND LOW NOISE AMPLIFIERS

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

NARROW BAND LOW NOISE AND MEDIUM POWER AMPLIFIERS

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

ULTRA-BROADBAND & MULTI-OCTAVE BAND AMPLIFIERS

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

LIMITING AMPLIFIERS

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

AMPLIFIERS WITH INTEGRATED GAIN ATTENUATION

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

LOW FREQUENCY AMPLIFIERS

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

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Dan Massé, Associate Technical Editor

Lockheed Martin Awarded Contract to Defend Against Anti-ship and Missile Threats

The US Navy awarded Lockheed Martin a $119 M contract to demonstrate its conceptual designs for the Air and Missile Defense Radar (AMDR) intended to upgrade the Navy's future capability against advanced anti-ship and ballistic missile threats. Lockheed Martin was one of three industry teams competitively selected to advance its designs for the AMDR S-band radar and the radar suite controller under a 24-month technology demonstration phase contract.

"Lockheed Martin's AMDR design provides a high performance radar suite capable of supporting simultaneous multi-mission operations in stressing environments, balanced with affordable lifecycle costs," said Carl Bannar, Vice President of Radar Systems. "Our team has developed an open-architecture and modular design that has matured technologies through significant investment by our company and the Navy."

The AMDR suite is scalable and can accommodate current and future mission requirements for multiple platforms. The AMDR S-band radar will provide volume search, tracking, ballistic missile defense discrimination and missile communications. The AMDR X-band radar, part of a separate procurement, will provide horizon search, precision tracking, missile communication and terminal illumination. Both radars will be integrated by the radar suite controller, which will provide the appropriate interfaces between the two radar bands and the ship's combat system.

Raytheon to Design Technology Demonstrator for Air and Missile Defense Radar

Raytheon Co. also received a $112.3 M contract to design and develop an S-band radar and radar suite controller technology demonstrator for the US Navy's new Air and Missile Defense Radar (AMDR). This highly advanced system will provide unprecedented capabilities for the Navy's Arleigh Burke-class destroyers. AMDR is being developed to fill capability gaps identified by the Joint Requirements Oversight Council in May 2006. The radar suite will consist of an S-band radar, X-band radar and radar suite controller. The system is scalable, enabling installation and integration with multiple platforms to meet the Navy's current and future mission requirements.

The radar suite will consist of an S-band radar, X-band radar and radar suite controller.

The AMDR S-band component will provide volume search, tracking, ballistic missile defense discrimination and missile communications. Raytheon's skills and expertise in dual band radar technology will ensure that AMDR's S- and X-band radars operate in coordination across a variety of operational environments.

Raytheon recently validated the maturity of its S-band capability with a demonstration of the system's advanced power amplifier and beamforming technology. System reliability and performance were confirmed, propelling the company closer to a full demonstration of a highly capable, low-risk solution for the Navy.

Raytheon has partnered with General Dynamics Advanced Information Systems and shipbuilder Gibbs & Cox in the concept development of this next-generation radar solution.

Under the contract, Raytheon IDS will design and develop radar arrays for far-field testing. The technology demonstrator will reduce risk and refine the system concept in preparation for the engineering and manufacturing development phase of the program. Work on the program will be performed at Raytheon IDS Headquarters, Tewksbury, MA; at the Surveillance and Sensors Center, Sudbury, MA; at the Seapower Capability Center, Portsmouth RI; and at the Integrated Air Defense Center, Andover, MA.

Northrop Grumman Awarded Contract to Develop AMDR Phase II Concepts

The US Navy has also awarded Northrop Grumman Corp. a $120 M contract to conduct technology development for the Air and Missile Defense Radar (AMDR). A next generation radar system planned for the Navy's Future Surface Combatant, AMDR is designed as a scalable, multi-mission radar system. AMDR is comprised of an X-band and an S-band radar along with a radar suite controller and is intended to provide unprecedented situational awareness to easily detect, track and engage ballistic missiles in high clutter environments.

"Under this two-year contract, Northrop Grumman will mature and demonstrate several key technologies required for the AMDR S-band radar and the radar suite controller," said Steve McCoy, Vice President of the Advanced Concepts business unit for the company's Electronic Systems sector.

Northrop Grumman is leveraging its extensive history of military S-band radar development along with modular, open architecture approach to provide a solution for AMDR that will scale to multiple ship classes and help protect the US Navy fleet for the next 40 years, McCoy added.

Northrop Grumman has proven active electronically scanned array technology developed for airborne and surface-based platforms and is a leading integrator of shipboard electronics. The corporation has delivered more than 500 S-band radar systems and is currently the supplier on a large S-band system for the Navy.

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Harris Receives $9.6 M to Deliver Type-1, Suite B Handheld Radios

Harris Corp., an international communications and information technology company, has received a $9.6 M order for the Harris RF-310M-HH, a new handheld tactical radio designed to simplify and improve secure communications interoperability for US and allied forces in combat. Harris is supplying RF-310M-HH multiband, multi-mission, software-defined radios under the US Department of Defense Coalition Readiness Support Program to provide secure communications interoperability to the International Security Assistance Force (ISAF) in Afghanistan. The RF-310M-HH is the first NSA Type-1, Suite-B certified tactical radio for protecting voice and data transmissions up to SECRET level classification.

This capability will enable secure communication between coalition units using RF-310M-HH radios and US forces using either RF-310M-HH or other Suite B tactical radio products, such as the Falcon III® AN/PRC152(C) handheld and AN/PRC-117G manpack radios. The AN/PRC-152 and AN/PRC-117G radios contain both US Type-1 Suite A and Suite B encryption, providing secure interoperability with the RF-310M-HH in support of various operational scenarios.

"The RF-310M-HH was developed in response to the increased use of multi-national coalitions on today's battlefield," said Steve Marschilok, President, Department of Defense business, Harris RF Communications. "Nations in these coalitions utilize many different types of radios, each with their own sovereign encryption technologies, making inter-coalition communication a challenge."

The RF-310M-HH radio uses the Sierra IIB programmable encryption module, certified by the NSA in June 2009. In addition to allowing secure interoperation with other tactical radios, the RF-310M-HH also hosts the APCO P25 waveform. The P25 waveform provides interoperability with radios used by police and other emergency response organizations, allowing the RF-310M-HH to span the gap between tactical and public safety networks. With its programmable encryption technology, Software Communications Architecture (SCA) and coverage of the entire 30 to 512 MHz frequency range, the RF-310M-HH is the ideal solution to interoperability challenges presented in coalition operations.

The RF-310M-HH is the first NSA Type-1, Suite-B certified tactical radio for protecting voice and data transmissions up to SECRET level classification.

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Indra to Design Radar System for ESA

The European Space Agency (ESA) has awarded Indra three contracts worth a total of €5.4 M to design a future radar system for the surveillance of objects close to the Earth as well as the development of its working prototype. The company will also determine its location and develop the simulator for the system. The contracts strengthen Indra's role within the preparatory stage of the Space Situational Awareness (SSA), promoted by ESA.

The working prototype will prove, in a real scenario, its capacity to detect objects in Low Earth Orbits at altitudes between 200 and 2,000 km and will require Indra to design one of the most advanced radar systems capable of operating at such distances. The study of feasibility and the tests will last for two years—until 2012. The results will form the basis for ESA to lay the basis for building the actual detection network. The tests with the prototype will help determine the structure and amount of equipment so that costs can be estimated.

The project is within the strategy of the European Union to gain further insight into the space environment...

The other two contracts include the development of a radar simulator that will predict the radar's system performance and will be used to optimise its final design. Indra is also considering the locations for the optimisation of the future surveillance radar in Europe and the report will help ESA decide the location.

The initiative seeks to increase security of space missions in the event of collisions with orbiting objects, magnetic storms or meteors. The project is within the strategy of the European Union to gain further insight into the space environment by proprietary means, which will undertake surveillance of nearby orbits and will monitor more distant zones. For this purpose, different technologies such as radars and optic telescopes, etc., will be employed.

USA ETSI Meeting Stresses Global Importance of M2M

The global significance of machine-to-machine (M2M) communications was demonstrated by the first meeting outside Europe of the M2M technical committee of the European Telecommunications Standards Institute (ETSI). The meeting of the committee, which defines the operation of future M2M communications and the efficient integration onto wireless networks, was held in Philadelphia, PA, and hosted by InterDigital Inc.

Over 40 telecommunications experts from leading companies and organizations around the world participated in shaping wireless technologies that connect people and devices—ranging from remote monitoring and control to e-health, smart transportation and intelligent homes—changing the way people live, work and play.

As an adjunct to the meeting, the host held a successful demonstration of a working prototype of an end-to-end architecture that interconnects cellular and non-cellular (e.g. WPAN, WLAN) networks seamlessly, using the cellular network to remotely manage and control both cellular and non-cellular M2M devices. This architecture is being standardized within ETSI in order to enable a broad range of M2M applications in vertical markets such as smart energy, healthcare, transportation and security.

... architecture is being standardized within ETSI in order to enable a broad range of M2M applications in vertical markets...

France Telecom's Marylin Arndt, Chairman of the ETSI M2M committee, announced that the first step of the standardization process, defining requirements for the system, had been successfully completed and that work continues to develop the detailed standards. A first release of these is foreseen for the first semester of 2011, enabling the global industry to deliver M2M services.

Thales Constructs Air Traffic Management Centre in Melbourne

Thales Australia has begun the construction of a new research and development centre in Melbourne, Victoria, Australia, aimed at shaping the future of Air Traffic Management (ATM). The Centre for Advanced Studies in ATM (CASIA) will be at the forefront of the company's work on new ATM offerings for Australia and export markets around the world. Beyond cutting-edge R&D, CASIA will be available for use by Thales' customers, partners and suppliers. The company will also strengthen its ties with Victorian universities and other educational institutions through shared research initiatives.

"Thales Australia is ideally placed to offer local, regional and global customers the most innovative ATM solutions..."

The centre will be located within Thales Australia's existing premises at Melbourne's World Trade Centre, and will begin operations early in 2011. It will pioneer innovative solutions and explore new technologies that will directly contribute to making air travel safer, more efficient and greener, while delivering a host of other benefits to local and international Air Navigation Service Providers.

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Chris Jenkins, Thales Australia's CEO, said CASIA represented the next chapter of ATM in Australia. "CASIA is the result of long term investment in our ATM business, which has grown from just a few employees 15 years ago to a global centre of excellence employing 400 people in highly skilled jobs."

"Thales Australia is ideally placed to offer local, regional and global customers the most innovative ATM solutions in the world today, building on our success with the Eurocat system and enabling effective airspace management in an era of increasing air traffic and technological complexity."

ip.access Signs Femtocell Agreement with Qualcomm

ip.access has signed a femtocell technology development agreement with Qualcomm Inc. that will enable ip.access to develop W-CDMA residential and enterprise femtocell products using Qualcomm's Femtocell Station Modem™ (FSM™) chipset platform. The company is also a licensee of Qualcomm's femtocell patent licensing program.

Femtocells are low power access points that create a mobile phone signal in homes, offices, shops and other locations that are hard to reach using a regular outdoor mobile phone network. The agreement between ip.access and Qualcomm comes at a time when network operators in North America, Europe and Asia have been accelerating their femtocell deployments.

"We are now seeing a mature femtocell ecosystem with diversity of supply for all of the key technology components."

"Our FSM chipset provides a highly integrated solution including radio frequency capabilities, baseband, network listen, GPS and a 1 GHz Snapdragon™-based processor," said Ed Knapp, Senior Vice President of Marketing at Qualcomm Flarion Technologies. "The FSM reduces the cost of a femtocell access point while providing best in class interference management techniques and other advancements. ip.access has established a leading position in this exciting new technology area, and we are very pleased to be collaborating with them."

"Femtocells and picocells are becoming an important part of today's 3G networks, and will be increasingly used as part of future network evolutions," said ip.access CEO Stephen Mallinson. "We are now seeing a mature femtocell ecosystem with diversity of supply for all of the key technology components. Qualcomm's expertise, industry knowledge and connections make them a valuable potential supplier for ip.access."

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TYPICAL SPECIFICATIONS

MODEL	FREQ. (GHz)	GAIN (dB)	POUT (dBm) @ 1 dB Comp.	NOISE FIG. (dB)	PRICE (1-9)
ZVA-183X+	0.7-18	26	+24	3.0	845.00
ZVA-213X+	0.8-21	26	+24	3.0	945.00

Note: Alternative heat-sink must be provided to limit maximum base plate temperature.

MODEL	FREQ. (GHz)	GAIN (dB)	POUT (dBm) @ 1 dB Comp.	NOISE FIG. (dB)	PRICE (1-9)
ZVA-183+	0.7-18	26	+24	3.0	895.00
ZVA-213+	0.8-21	26	+24	3.0	995.00

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Dan Massé, Associate Technical Editor

GPS, Accelerometers and Gyroscopes Will Add Functions to Smartphones

ABI Research anticipates that the sensor-driven user interface (UI) will be an emergent theme in the next wave of mobile UI innovation, turning objects, locations and people into networked, interactive elements. By 2013, 85 percent of smartphones will ship with GPS, over 50 percent will ship with accelerometers and almost 50 percent will have gyroscopes.

"The growth of sensors in smartphones will be driven by applications such as gaming, location awareness and augmented reality, as well as the expansion of motion-based commands," says Senior Analyst Victoria Fodale. "The high-level operating system of a smartphone, which provides open application programming interfaces (API), has facilitated the use of data from cameras, sensors and GPS receivers."

When an accelerometer is combined with a gyroscope, developers are able to create applications that can sense motion on six axes: up and down, left and right, forward and backwards, as well as roll, pitch and yaw rotations. This gives a mobile device similar functionality to a game controller such as the Nintendo Wii. Prompted by Apple's UI innovations with the iPhone, smartphone OEMs have poured resources into UI design and development. Many OEMs, particularly those using Google's Android OS, developed their own custom UI overlays. Sensors will also help OEMs to innovate beyond a touchscreen UI for differentiation in the marketplace. However, added functionality must be balanced with ease of use. "There is an inherent paradox with technology," says Fodale. "As mobile devices integrate more technology, the UI must be kept simple enough to be intuitive for the user."

By 2013, 85 percent of smartphones will ship with GPS, over 50 percent will ship with accelerometers and almost 50 percent will have gyroscopes.

A new ABI Research study, "Mobile Device User Interfaces," examines the key market and technology trends for mobile device UIs. It highlights the UIs of leading device OEMs, and examines UI technologies including frameworks, touchscreens, GPS, sensors and other components. Forecasts include smartphones shipments by OS and region and mobile augmented reality application downloads, as well as the attach rate for key sensor and other components.

Global M2M Module Market Forecast to Reach $3.8 B

The Cellular Machine-to-Machine (M2M) communication market has been a challenging place for cellular embedded module vendors over the past 18 months. 2007 was this market's last "good year": 2008 saw both shipment volumes and revenue decrease. In 2009, the cellular M2M module market grew somewhat in terms of unit shipment volume, although rapidly declining module prices meant that 2009 revenue was still below that of 2007.

Despite its difficulties, however, this market is on an upward path. Unit shipment volume growth and the growing importance of 3G are already resulting in stronger performance and a new ABI Research study estimates that the market will be worth $3.8 B in 2015.

What were the cellular M2M module market size and module vendor market shares in 2009?

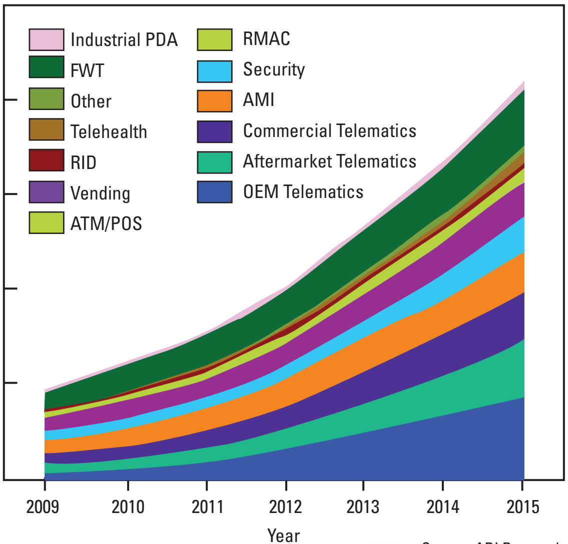
What is the cellular M2M module market forecast, by shipment and revenue, segmented by region, application, and air interface standard?

What are the strategic responses of cellular M2M module vendors to market forces and resulting market trends?

To learn more about the cellular M2M market and how it may affect business models now and in the future, please visit the ABI Research study "Cellular Machine-to-Machine (M2M) Markets," which discusses these trends, analyzes cellular M2M module vendor strategic responses, provides vendor market share data for the period 2003-2009, and forecasts cellular M2M module shipment and revenue growth for the period from 2007 through 2015, segmented by region, application and air interface standard.

It is part of the firm's M2M Research Service, which also includes other Research Reports, Research Briefs, Market Data, ABI Insights and analyst inquiry support.

Total Cellular M2M Module Shipments by Application World Market, OEM-Basis Forecast: 2009 to 2015



Source: ABI Research

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Almost 13 Million Fixed Wireless Terminals and Cellular Routers Will Ship in 2015

In 2015, shipments of fixed wireless terminals (FWT) and cellular routers will total nearly 13 million, according to the latest forecasts from ABI Research. The research includes market analysis for industrial terminals, business gateways, telephony adaptors—all relatively mature markets showing stable modest growth—and the newest market segment, mobile broadband routers, which will contribute the greatest increase in shipments.

"While they share underlying technologies, these devices/applications perform different roles and are used in diverse environments," says M2M Practice Director Sam Lucero. "Industrial terminals, as machine-to-machine devices, benefit from the growing business and government interest in telemetry and telematics. Business gateways now offer viable alternatives to DSL-based services and are increasingly used in remote branch offices. Telephony adaptors connect local analog voice phone systems to the world via a cellular network; the Chinese government is particularly interested in their use to extend telephony services to rural areas."

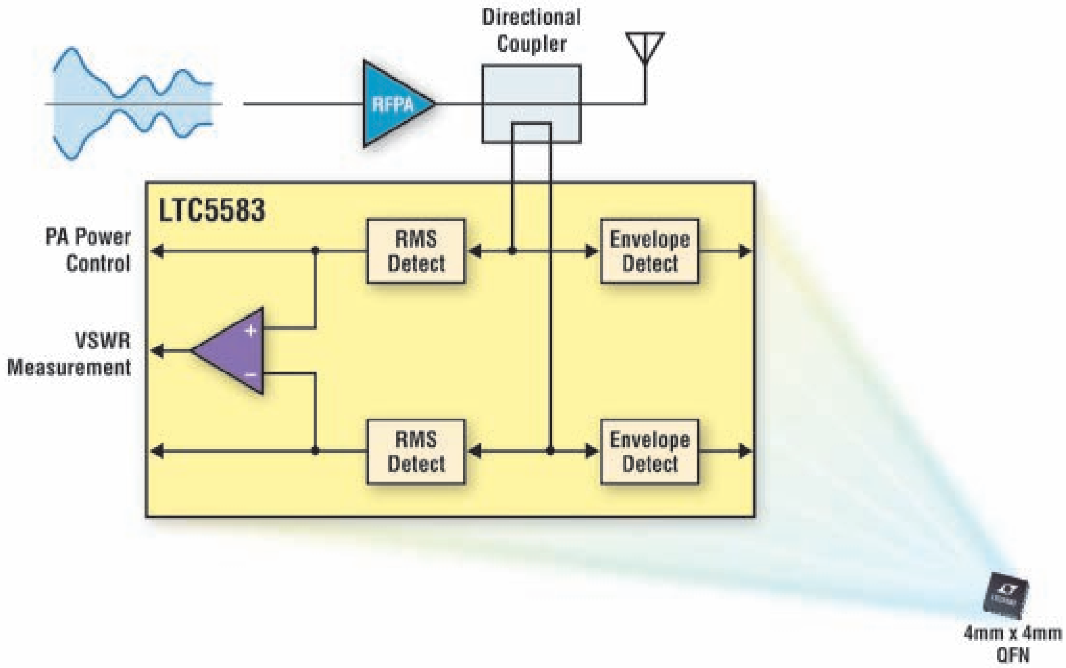
Principal Analyst Jeff Orr adds: "Mobile broadband routers allowing multiple devices to connect from 'anywhere' (a car, a hotel, a construction site) may be found in both consumer and business contexts. Businesses are already familiar with wired business continuity and redundancy methods. Now devices have 'grown wings,' becoming mobile broadband routers. This segment is starting from practically zero, and its top is not yet in sight."

"Industrial terminals, as machine-to-machine devices, benefit from the growing business and government interest in telemetry and telematics."

To make the most of this market, Orr and Lucero recommend that:

- Cellular embedded module vendors should seek new application opportunities in other FWT and M2M segments.
- Business gateway vendors should focus on data connectivity rather than voice connectivity.
- Developers of mobile broadband hotspot routers must understand early-adopter markets.
- Industrial terminal vendors should offer platforms that include cloud computing software delivery models.
- Mobile network operators should reconsider how pervasive Wi-Fi technology enables new subscribers and welcomes multi-device users.
- FWT and cellular broadband router vendors should consider entry into the business gateway market.

±0.5dB RMS Power Accuracy



6GHz Dual, Matched RMS Detector with Superior Isolation Simplifies Calibration

The LTC®5583 delivers outstanding channel-to-channel isolation and 60dB dynamic range using single-ended inputs, eliminating the need for external balun transformers. This dual RF detector offers exceptional accuracy of ±0.5dB error over a temperature range from –40°C to 85°C. The device provides a drop-in solution that simplifies design, improves accuracy and reduces costs.

▼ High Performance RMS Detectors

	LTC5583	LTC5582	LT®5581	LTC5587
Operating Frequency	40MHz to 6GHz	40MHz to 10GHz	10MHz to 6GHz	10MHz to 6GHz
# Channels	2	1	1	1
Dynamic Range (dB)	60	57	40	40
Detection Range (dBm)	–58 to 2	–56 to 1	–34 to 6	–34 to 6
Measurement Accuracy (–40°C to 85°C)	±0.5dB	±0.5dB	±1dB	±1dB
Output Interface	Log-Linear Voltage	Log-Linear Voltage	Log-Linear Voltage	12-Bit Serial ADC
Power Supply	3.3V/90mA	3.3V/42mA	3V/1.4mA	3.3V/3mA

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LINEAR TECHNOLOGY

Around the Circuit

Jennifer DiMarco, Staff Editor

INDUSTRY NEWS

L-3 Communications announced that it acquired **3Di Technologies LLC**. 3Di provides highly specialized end-to-end secure satellite communications utilized by forward-deployed United States special operations and in-theater personnel. The terms of the transaction were not disclosed. The business will be included in the Microwave Group within L-3's Electronic Systems segment.

AML Communications Inc., a designer, manufacturer and marketer of microelectronic assemblies for the defense industry, announced the launch of **Cal Mimix Inc.**, a wholly owned subsidiary. Cal Mimix has been in formation for the last six months as a fabless designer and developer of RF and microwave semiconductor products to serve commercial and military applications. The company has entered into manufacturing contracts with an on-shore foundry for production of GaAs-based semiconductor devices and has taken delivery of low rate production units.

Ceragon Networks Ltd., a provider of high-capacity, 4G/LTE-Ready wireless backhaul networks, announced that it has acquired all the shares of privately-held **Elxys Innovations**, Athens, Greece. Elxys is a designer of advanced, next-generation radio-frequency integrated circuits (RFIC). The acquisition solidifies Ceragon's leadership in RF design and allows it to further optimize its core radio technology. By strengthening the radio design team with a group of seasoned professionals in the RFIC domain, Ceragon will continue to offer its customers superior radio performance, higher capacities, reduced power consumption and better overall cost.

Rohde & Schwarz has entered into a cooperative agreement with the Belgian company **NMDG**, a highly specialized provider of nonlinear network analysis solutions that will enable Rohde & Schwarz to expand its extensive network analysis portfolio of nonlinear measurement solutions. Today's advanced RF components and systems need to support complex modulation techniques and accommodate even wider transmission bandwidths. The goal is to achieve higher data throughput while simultaneously lowering power consumption.

W.L. Gore & Associates Inc. has entered into a strategic partnership with **A.E. Petsche Co.** to provide GORE® FireWire® Cable Products for the Joint Strike Fighter (JSF) F-35 Lightning program, the Department of Defense's next-generation strike aircraft system for the Navy, Air Force and Marines. A.E. Petsche will manage the requirements of all program partners to ensure optimum stocking levels and timely distribution.

Agilent Technologies Inc. announced that its application for a seat on the board of directors was accepted by the PXI Systems Alliance Board at its annual meeting on September 15. Following the recent announcement of 46 new PXI and AXIe modular products, Agilent applied for sponsor-level membership, the highest level of membership within the PXI Systems Alliance. The primary goal of the PXISA is to improve the effectiveness of CompactPCI-based solutions in measurement and automation through use of the PXI specification.

Linwave Electronic Manufacturing Services, a division of Linwave Technology, announced the opening of a new UK-based high technology microwave manufacturing centre. The facility will support key activities of diode manufacturing, MMIC packaging and high frequency/precision manufacturing solutions. The technology partnership between **e2v** and Linwave Technology has been enabled by the transfer of equipment, processes and 15 highly skilled personnel into the new facility.

ON Semiconductor, a supplier of high performance, energy efficient silicon solutions for green electronics, recently announced plans to expand production capacity at its 6-inch (150 mm) wafer manufacturing facility in Oudenaarde, Belgium, by approximately 40 percent with a total equipment investment in 2010 of 12.3 M EUR (or $15.78 M USD). The Belgium factory specializes in the manufacture of application specific high-voltage technologies for the automotive and industrial industry, and in integrated and discrete standard products for a wide range of market segments.

East Coast Microwave is pleased to announce the company's recent ISO 9001:2008 certification. Certificates available upon request. As of October 1, 2010, East Coast has moved to an expanded facility. The new address: 70 Tower Office Park, Woburn, MA 01801.

Emerson Network Power Connectivity Solutions, a business of Emerson and a leader in enabling Business-Critical Continuity™, announced that it has been awarded the **Digi-Key** Superior Sales Performance by Superior Engagement Award.

CONTRACTS

SenarioTek LLC, a provider of custom RF/microwave test and measurement solutions, has been awarded a contract by **Orbital Sciences Corp.**, Dulles, VA, to provide test technology to extend the capabilities of Orbital's installed payload test systems for the HYLAS 2 satellite payload. The SenarioTek solution will provide Orbital the highest level of RF accuracy and performance and reduces the test program risk by better characterizing the uncertainties introduced by the test system.

VTI Instruments announced that it has been selected by

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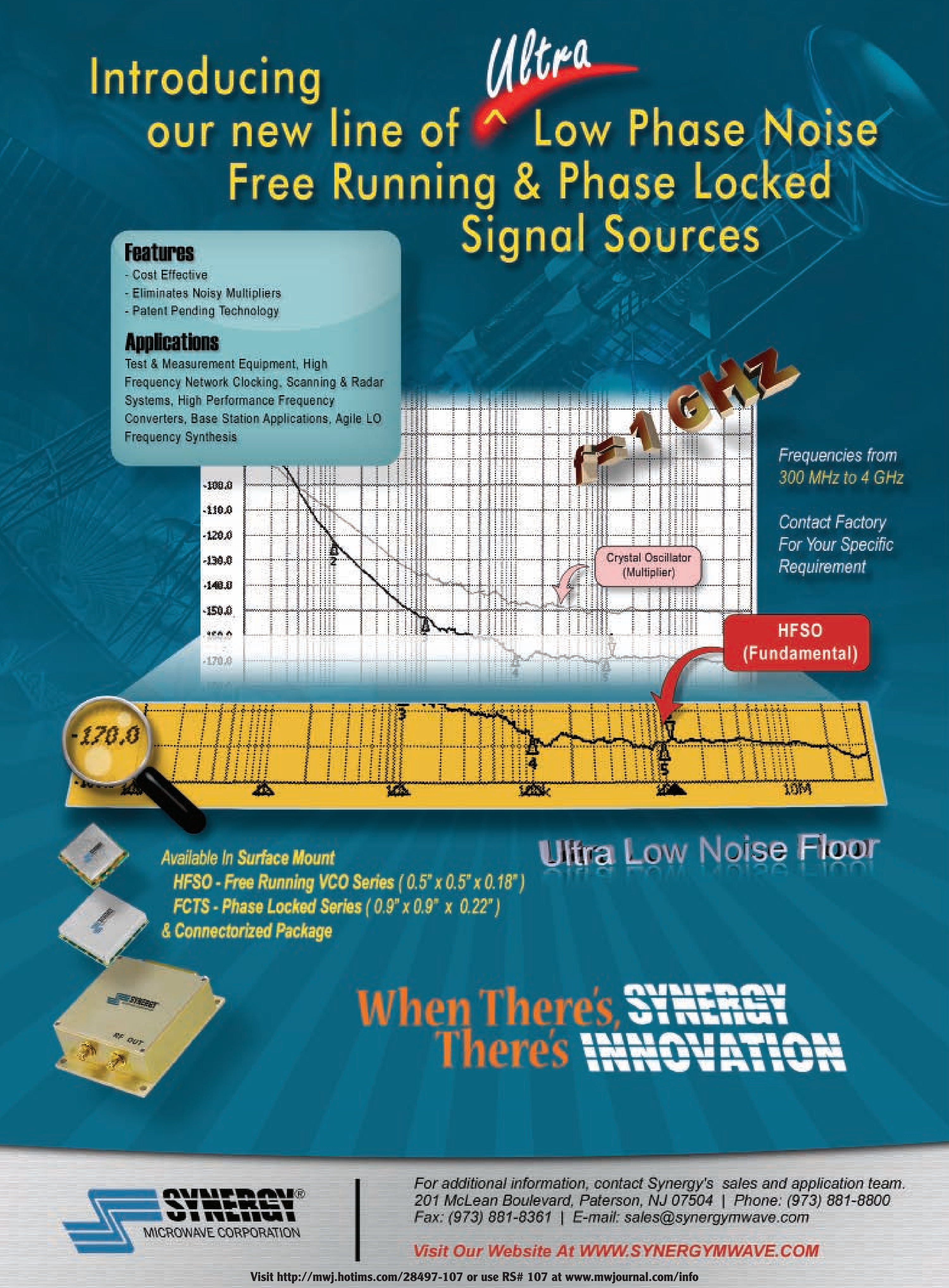
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-110.0
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10M
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Lockheed Martin to provide the Enhanced Switching and General Purpose Interface (eGPI) subsystem for the US Navy's eCASS (Electronic Consolidated Automated Support System) Program. eCASS will replace CASS as the US Navy's core test system and ensures that fleet aircraft can maintain mission-ready status by facilitating quick diagnostic and repair of line-replaceable subassemblies.

WiSpry Inc., a leader in tunable RF semiconductor products for the wireless industry, announced that it has won, along with key partners, a four year, $48 M DKK ($8 M USD), Smart Antenna Front End (SAFE) project to be funded by Denmark's High Technology Foundation. WiSpry will work together with Aalborg University (AAU), antenna specialist Molex Interconnect, and chipset leader Infineon Technologies to develop tunable antennas and RF front-ends based on WiSpry's tunable RF technology. To work closely with its partners, WiSpry will be establishing a technical presence in Aalborg, Denmark. The Smart Antenna Front End (SAFE) project's goal is to develop antennas and front-end technology platforms for the next generation of mobile handsets and media devices. SAFE will enable significant reduction in the size and cost of mobile devices while increasing their overall efficiency.

NEW MARKET ENTRY

Wireless Applications Corp., a Bellevue, Washington-based Telecom Infrastructure software and services company, announced that it has beta-released a suite of web service Application Programming Interface (API) for performing telecom infrastructure analysis called the Telecom API Suite. Carriers, tower companies, web app programmers and wireless service providers alike can immediately begin amping up their software with power from the Wireless Applications' cloud service.

PERSONNEL


▲ *Bob Tavares*

Crane Aerospace & Electronics, a segment of Crane Co., has announced the appointment of **Bob Tavares** as Vice President, Microwave Solutions for the Electronics Group. Tavares will lead the Microwave Solutions business. Tavares comes to Crane Aerospace & Electronics with an extensive background in the microwave industry. He has spent most of his career at Tyco Electronics, M/A-COM Division. He started his career as an engineer and then progressively advanced his career to his latest position of Vice President, General Manager, where he was responsible for setting the strategic direction, growth and profitability of a $320 M RF and microwave multi-site business, making a diverse set of highly custom and application specific products. He also created divisional LEAN and 6 Sigma programs, which became the foundation for the Operational Excellence program in M/A-COM.

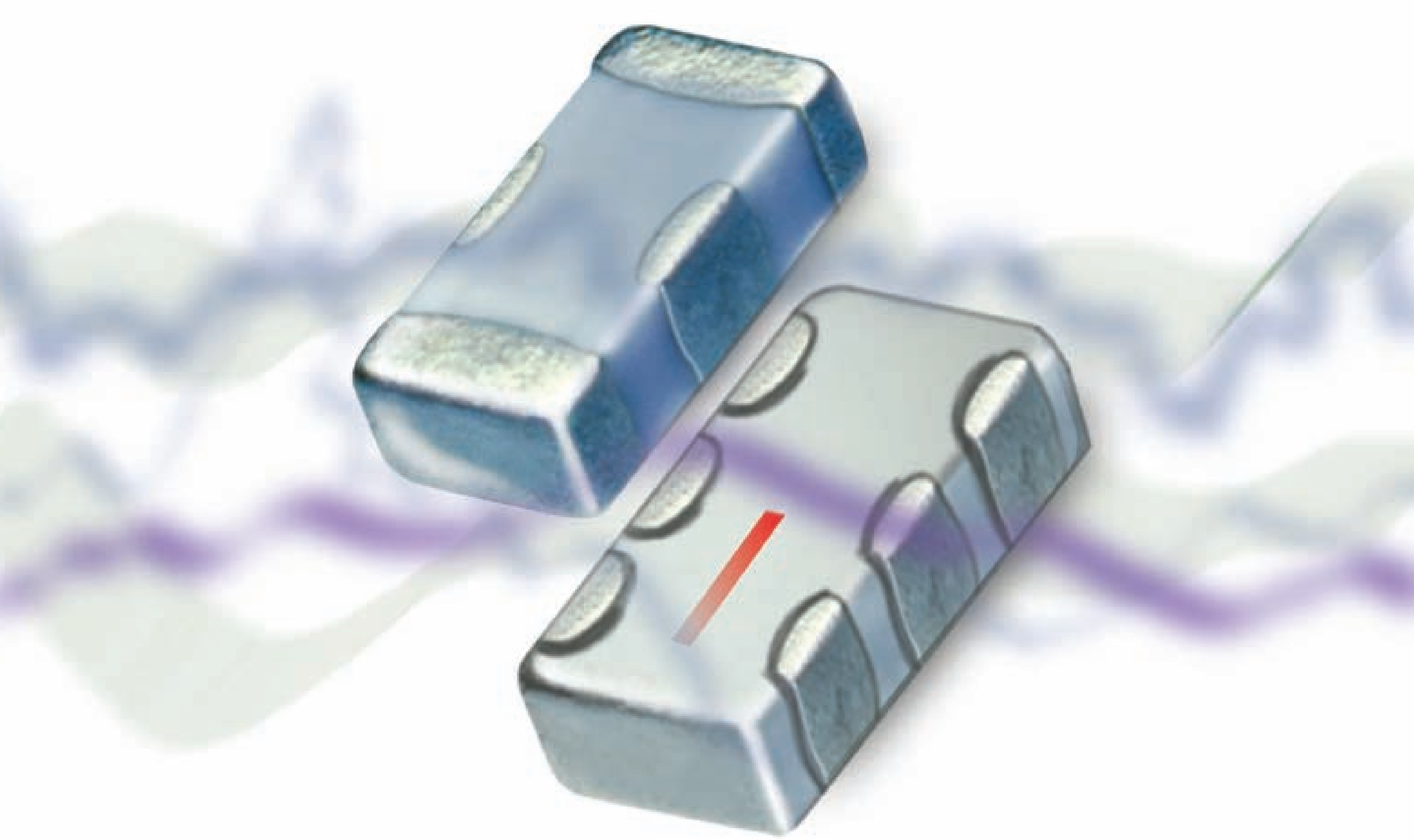
REP APPOINTMENTS

Anritsu announced the appointment of **Trescal** as its sole certified service partner in France. This certification means that Trescal will be providing OEM backed calibration and repair services on Anritsu products via their network of labs throughout the country. Anritsu will be directly supporting Trescal, as an integral part of this relationship, to ensure the continuing high quality service support of Anritsu products for its French customers.

Anatech Electronics announced three manufacturer's representatives each responsible for sales of the company's RF and microwave filter and passive products in different areas of the US. **FCS Group** will represent Anatech Electronics in Colorado, Montana, Utah and Wyoming. The company has multiple sales offices in Colorado and more than 15 years of experience in technical sales and support, along with contacts in key aerospace, communications and electronics manufacturing companies. The FCS Group can be reached at (303) 587-3353. **Martronix** will represent Anatech Electronics in Southern California and southern Nevada. The company has two offices based in the Southern California and has been representing high-tech companies for more than 20 years. They can be reached at (805) 499-6385. **Electronic Instrument Associates** will represent Anatech Electronics in Illinois, Indiana, Iowa, Michigan, Minnesota, North and South Dakota, and Wisconsin. The company was founded in 1971 and covers the Midwest through three offices in Chicago, Minneapolis and Indianapolis. Electronic Instrument Associates can be reached at (630) 924-1600.

MITEQ Inc. announced the appointment of **Test Midwest LLC** as the company's exclusive sales representative in Illinois, Iowa, Kansas, Minnesota, Missouri, Nebraska, North and South Dakota, and Wisconsin. Test Midwest LLC will represent MITEQ's Component division of products, which includes amplifiers, mixers, frequency multipliers, passive power components, switches, attenuators, limiters, phase shifters, IF signal processing components, oscillators, synthesizers, integrated multifunction assemblies and fiber optic products. Test Midwest LLC can be contacted at (866) 639-3311.

International Manufacturing Services Inc. (IMS), a manufacturer and supplier of high quality thick and thin film resistors, terminations, attenuators, planar dividers, planar filters and thermal management devices to the electronics industry, announced the appointment of **E.G. Holmes** as its Dixie States representative. E.G. Holmes and Associates Inc. was organized in 1955 by Edward Holmes and remains focused on sales of RF, microwave and fiber optic/electro-optic components, and subsystems. Sales territories encompassed by EGH include South Carolina, Tennessee, Mississippi, Alabama, Georgia and Florida.



A 470 TO 960 MHz RESONANT ANTENNA: COVERING UHF MOBILE TV AND CDMA/GSM WITHOUT TUNING CIRCUITS

A novel wideband antenna has been developed for UHF mobile digital TV. The new antenna has a bandwidth of more than 68 percent. It is an unbalanced resonant antenna that does not need any tuning or matching circuits. It resonates from 470 to 960 MHz and can cover all bands of UHF mobile digital TV (470 to 862 MHz or a portion of it), CDMA/TDMA/GSM800 (824 to 894 MHz) and E-GSM900 (880 to 960 MHz). The overall size of the new antenna is very small and its manufacturing costs are very low. It has a very small cross-section and is made of a flexible material. Bending the antenna in more than one direction considerably reduces the effect of the human body and the surrounding environment on the antenna. It also increases its sensitivity to different polarizations. Since the new antenna is multi-polarized, it significantly reduces the need for separate diversity antennas. The new antenna can be used as an internal, external or partially internal and partially external antenna.

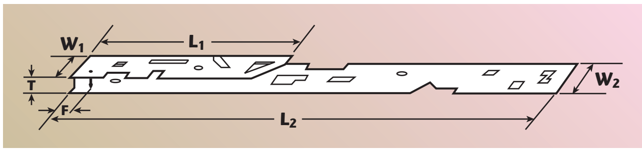
Antennas for mobile TV are usually required to have a very wide frequency band. For example, ultra high frequency digital video broadcasting-handheld (UHF DVB-H) is designed to work in the frequency band from 470 to 862 MHz or a portion of it. This is a very wide bandwidth, which is difficult to cover with a single resonant antenna. Therefore, matching circuits are usually used to tune the antenna for this band[1] or even for narrower bands such as 470 to 770 MHz[2] and 470 to 702 MHz.[3] Matching circuits increase the complexity and the costs of the antenna and also reduce its efficiency.

On the other hand, the mobile TV band overlaps with the CDMA/GSM800 band (824 to 894 MHz). It is also too close to the E-GSM900 band (880 to 960 MHz). This overlapping may cause severe coupling and interference between antennas of both applications, especially if they are placed close to each other in the limited space inside handsets, which are getting smaller and smaller. A novel solution to

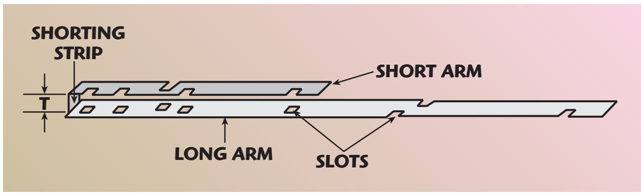
MOHAMED SANAD
Amant-Antennas, Giza, Egypt
NOHA HASSAN
Cairo University, Cairo, Egypt

overcome this interference problem is to use only one antenna that can cover all bands of UHF mobile digital TV, CDMA/TDMA/GSM800 and E-GSM900. In this case the antenna is required to cover an overall band from 470 to 960 MHz, which is very challenging.

A wideband antenna for UHF mobile digital TV has been developed. The new antenna can cover a bandwidth of more than 68 percent. It is an unbalanced resonant antenna that does not need any tuning matching circuits. It resonates from 470 to 960 MHz and hence can cover all bands of UHF mobile digital TV, CDMA/TDMA/GSM800 and E-GSM900. Of course, if it is undesirable to merge the frequency bands of mobile TV and CDMA/GSM, the new antenna can be designed with a narrower bandwidth that can only cover the mobile TV band or any other band. Actually, narrowing the bandwidth increases the efficiency of the antenna. The new antenna can be used with cellular phones, palmtop, notebook, laptop computers or any other portable communication equipment.



▲ *Fig. 1 Geometry of the new antenna.*



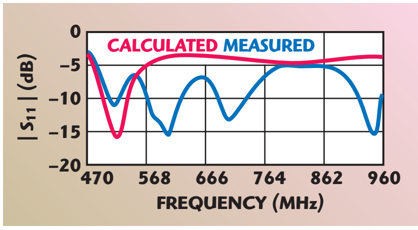
▲ *Fig. 2 The selected sample antenna configuration.*

DESCRIPTION OF THE NEW ANTENNA

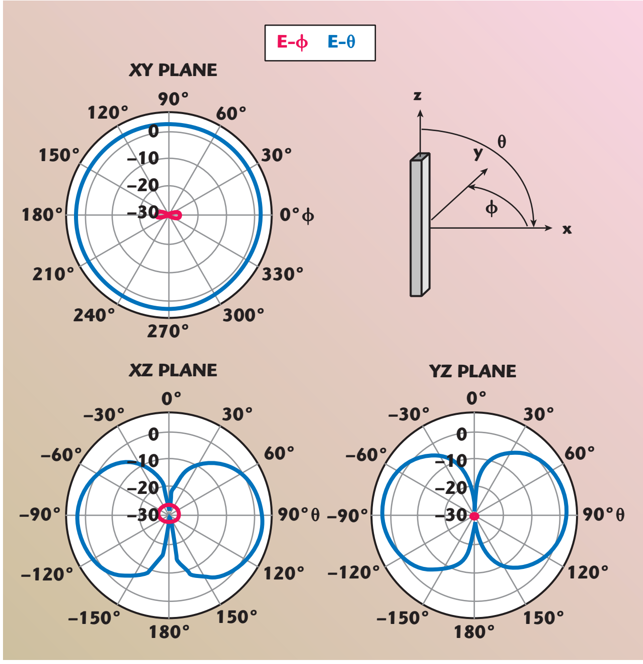
Figure 1 shows the geometry of the newly developed UHF digital mobile TV antenna. It consists of two narrow printed metallic arms connected together by a shorting metallic strip. The two arms may be parallel to each other or may have any angle between them. The length of the short arm is L_1 and its width is W_1 while the length of the long arm is L_2 and its width is W_2, the thickness (the height) is T and the antenna is fed at a distance F from the shorted edge. The two arms of the antenna can have equal or unequal widths W_1 and W_2. Furthermore, the two arms can be shaped in different ways in order to optimize the antenna performance.

As shown, each arm has a set of slots having different configurations. These slots can be circular, rectangular, square, triangular or other shapes. The arm lengths of the new antenna, especially the length of the short arm, are the main parameters that determine the operating frequency of the antenna. The feed location is adjusted in each configuration in order to improve the return loss as much as possible. The bandwidth, the peak gain and the efficiency of the antenna are mainly determined by the widths of the two arms, the angle between them, the thickness of the antenna and the configurations of the slots, which are all optimized together in order to enhance the antenna performance, especially the bandwidth.

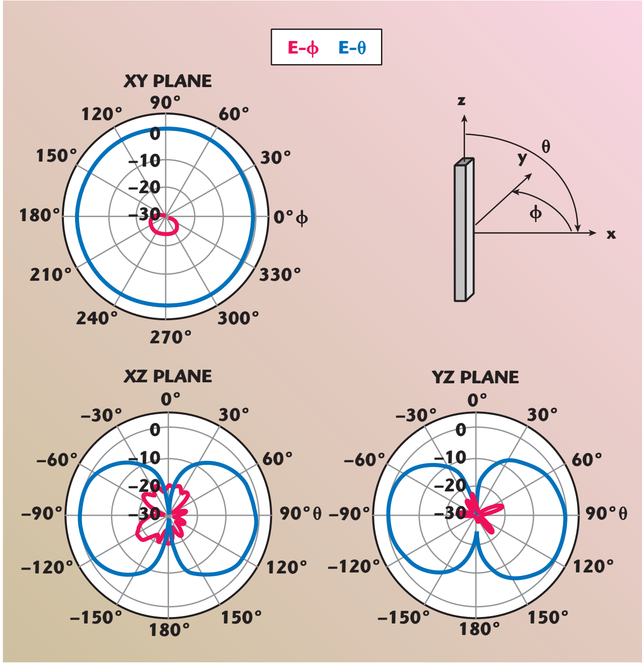
The antennas are completely self-contained and do not need extended ground planes or any additional components. Thus, the new antenna can be mounted anywhere, inside or outside any handset, because the antenna does not use a part of the handset as an extended ground plane, which usually happens with most available internal antennas. Furthermore, the antenna is made of a flexible printed material and can be bent and/or folded in different forms in order to fit any available space inside or outside the handset. Actually, it can be used as an internal, external or partially internal and partially external antenna. Moreover, the overall size of the antenna is small and its manufacturing costs are low.



▲ *Fig. 3 Calculated and measured return loss of the new antenna.*



▲ *Fig. 4 Calculated radiation patterns at 800 MHz.*



▲ *Fig. 5 Measured radiation patterns at 800 MHz.*

RESULTS

Different prototypes of the new mobile TV antenna have been designed, manufactured and tested. The results for a selected sample antenna configuration are presented. The sample is made of a flexible printed material "PET" with a dielectric constant $\varepsilon_r = 3.5$ and a loss tangent $\delta = 0.015$. The geometry of the selected antenna configuration is shown in **Figure 2**.

The two arms of the selected sample antenna are parallel to each other. The length L_1 of the short arm is 11.5 cm while the length L_2 of the long arm is 25 cm. The width W_1 of the short arm is 2.6 mm while the width W_2 of the long arm is 3.5 mm and the antenna thickness T is 2 mm. The overall size of the antenna is $25 \times 0.35 \times 0.2 = 1.75$ cm^3. It should be noted that this is the overall volume of the antenna because it does not require an additional ground plane, a matching circuit or any other components. All slots in both arms are selected to be rectangular in shape. The length of each slot is 5 mm and its width is 2 mm. The distance between the shorted edge and the first slot is D_1. The distances between the successive slots are $D_2, D_3, \ldots, D_8$, respectively.

The locations of the first five slots are similar in both arms. This means that the first five slots in the short arm are located exactly above the first five slots in the long arm. However, since the long arm is wider than the short arm, the first five slots are positioned close to the middle of the long arm, while they are located at the edge of the short arm as shown. The values of $D_1, D_2, \ldots, D_8$ are as follows: $D_1 = 5$ mm, $D_2 = 10$ mm, $D_3 = 15$ mm, $D_4 = 5$ mm, $D_5 = 45$ mm, $D_6 = 35$ mm, $D_7 = 5$ mm and $D_8 = 75$ mm.

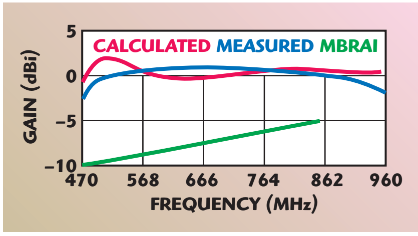
The return loss and the radiation

patterns of the antenna were numerically calculated using a software package that utilizes the moment method. They were also measured at the IMST antenna labs in Germany.[13] ***Figure 3*** shows the calculated and the measured return loss of the antenna. The measured return loss is better than 5 dB from approximately 470 to 960 MHz, which is more than 68 percent bandwidth. The calculated radiation patterns at 800 MHz, as a sample frequency, are shown in ***Figure 4***. The corresponding measured patterns are shown in ***Figure 5***.

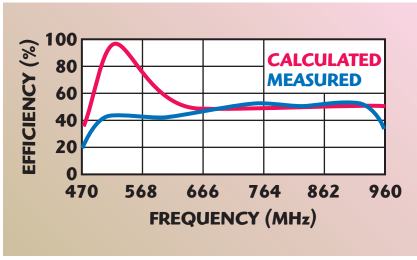
Figure 6 shows the calculated and the measured peak gain of the new antenna from 470 to 960 MHz, which is approximately 0 dBi over most of the band. The measured peak gain is much higher than the MBRAI specifications of UHF DVB-H mobile TV.[14] The calculated and measured efficiencies of the new antenna are shown in ***Figure 7*** from 470 to 960 MHz. The average measured efficiency over the whole frequency band is approximately 45 percent.

REDUCING THE ANTENNA SIZE

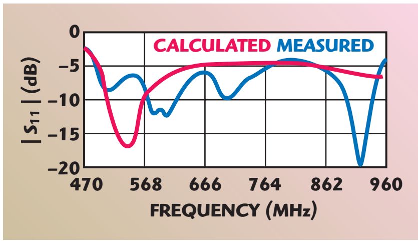
The antenna size can be further reduced by decreasing its width and/or its thickness. For example, if the width W_1 of the short arm of the above configuration is reduced from 2.6 to 2 mm while the width W_2 of the long arm is reduced from 3.5 to 2.8 mm and the antenna thickness T is also reduced from 2 to 1 mm, the overall size of the new antenna configuration is $25 \times 0.28 \times 0.1 = 0.7\ cm^3$. Thus, the size of the antenna is reduced from 1.75 to 0.7 cm^3. ***Figure 8*** shows the calculated and the measured return loss of the new antenna after reducing its overall size from 1.75 to 0.7 cm^3. The return loss is still approximately 5 dB over most of the band from 470 to 960 MHz. The calculated and the measured peak gain and efficiency of the antenna after reducing its overall size from 1.75 to 0.7 cm^3 are shown in ***Figures 9*** and ***10***, respectively. It is clear that the peak gain is still much



▲ *Fig. 6 Calculated and measured peak gain of the antenna.*



▲ *Fig. 7 Calculated and measured efficiency of the antenna.*



▲ *Fig. 8 Calculated and measured return loss of the reduced size antenna.*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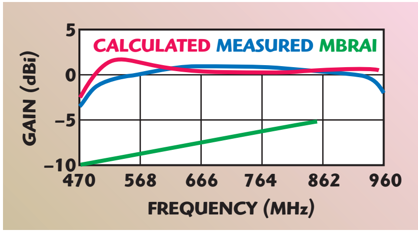
higher than the MBRAI specifications of UHF DVB-H mobile TV.

The width W_2 of the long arm of the new antenna is further reduced from 2.8 to 2 mm in order to make it equal to the width W_1 of the short arm. The overall size of the new antenna configuration is now 25 × 0.2 × 0.1 = 0.5 cm^3. ***Figure 11*** shows the calculated and the measured return loss of the new antenna after the second reduction in its size from 0.7 to 0.5 cm^3. The return loss is better than 6 dB over most of the band from 470 to 960 MHz. The measured peak gain of the 0.5 cm^3 antenna is still much higher than MBRAI specifications of UHF DVB-H mobile TV. It is not shown because of space limitations.

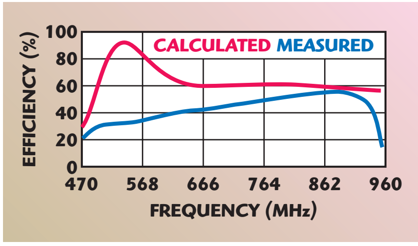
From all the results shown above, it can be seen that there are some considerable differences between the calculated and measured results. This is because the software package used assumes an infinite substrate. This assumption affects the accuracy of calculations when the width of the antenna substrate is very narrow as in the earlier antenna configurations. The difference between calculations and measurements is significant in efficiency curves because the efficiency depends on radiation patterns in all directions and in all planes; hence, there is an accumulated reduction in the accuracy of calculations. This was clearer at the lower part of the frequency range, where the antenna width is very narrow in terms of wavelengths. Furthermore, the difference between calculations and measurements was further increased when the width of the antenna was reduced from 3.5 to 2.8 mm, as can be observed by comparing Figures 7 and 10.

BENDING AND FOLDING THE ANTENNA

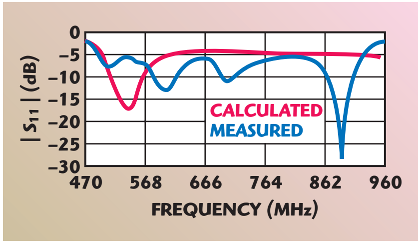
As mentioned before, the new antenna has a very small cross-section area and is made of a flexible printed material that can be easily bent and/or folded in order to fit the available space in any wireless equipment. Therefore, although the length of the



▲ *Fig. 9 Calculated and measured gain of the reduced size antenna.*



▲ *Fig. 10 Calculated and measured efficiency of the reduced size antenna.*



▲ *Fig. 11 Calculated and measured return loss of the 0.5 cm³ antenna.*

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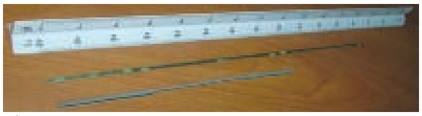

Fig. 12 *Straight and folded antennas.*

Fig. 13 *The antenna folded around a handset.*

antenna is 25 cm, it can be easily reduced in different ways. For example, the two ends of the antenna can be folded, as shown in ***Figure 12***, where the length is reduced from 25 to 16 cm without significant effect on its performance.

On the other hand, the new antenna can be folded to form a rectangular shape around the handset from the inside and/or outside, as shown in ***Figure 13***. The performance of the new antenna while it is folded in the form of a rectangular shape was also measured at IMST labs in Germany. The calculated and measured return loss of the folded antenna in the y-z plane is shown in ***Figure 14***. The calculated radiation patterns at a sample frequency 800 MHz are shown in ***Figure 15***; the correspond-

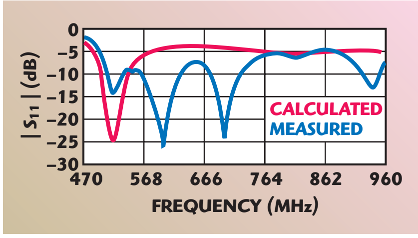

Fig. 14 *Calculated and return loss of a bent antenna.*

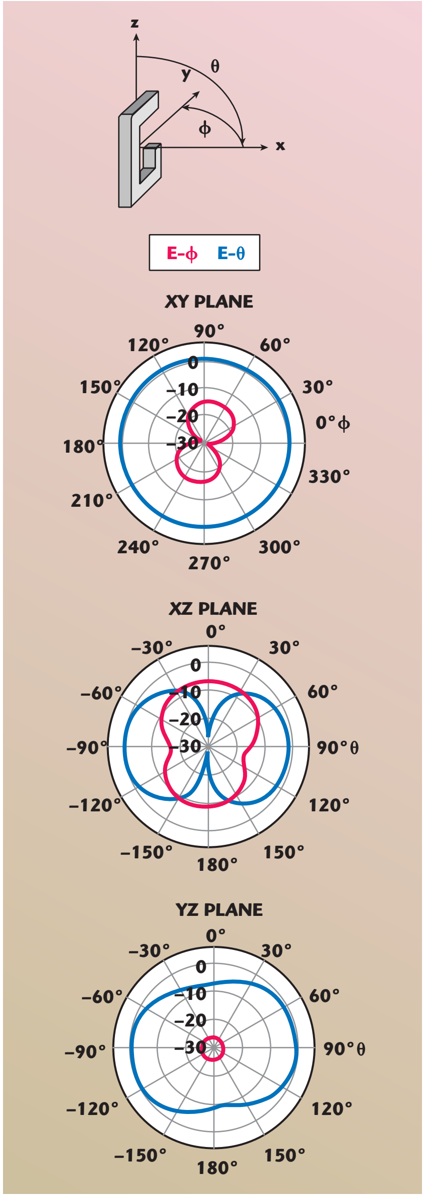

Fig. 15 *Calculated radiation patterns of the bent antenna.*


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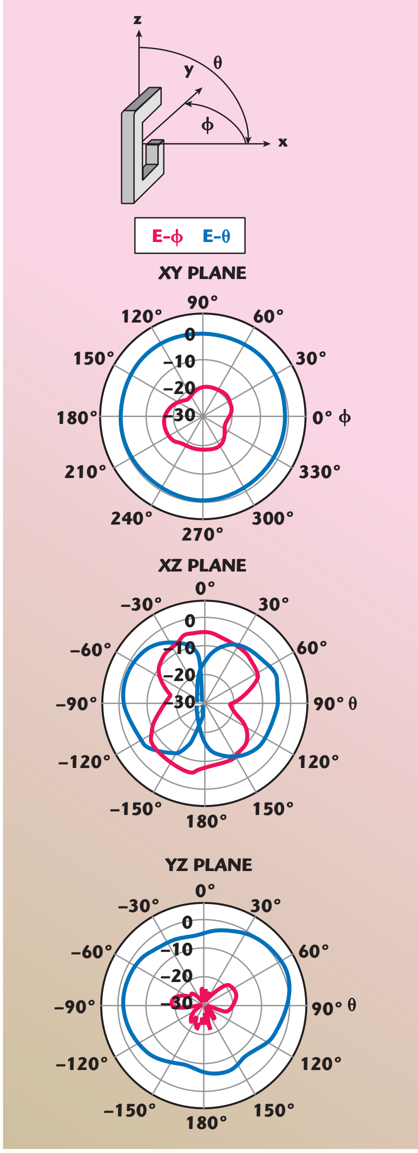
ing measured patterns are shown in ***Figure 16***. Comparing Figures 5 and 16, it can be seen that the main difference between the performance of the folded antenna and the straight antenna is the sensitivity to more than one polarization. It is clear that the straight antenna is sensitive to only one polarization. Folding the antenna in two perpendicular directions in the y-z plane makes the radiation patterns sensitive to two perpendicular polarizations. This is very important in MIMO and in all indoor applications, where the waves are randomly orientated because of multipath reflections and rotation of polarization.

MULTI-POLARIZED CONFIGURATIONS

From the above results, it was shown that bending the new antenna in two perpendicular directions significantly increased its sensitivity to two perpendicular polarizations. Of course, the optimum situation for polarization diversity is to make the antenna sensitive to three perpendicular polarizations. This can be achieved by bending the new antenna in three perpendicular directions. Bending the antenna in three

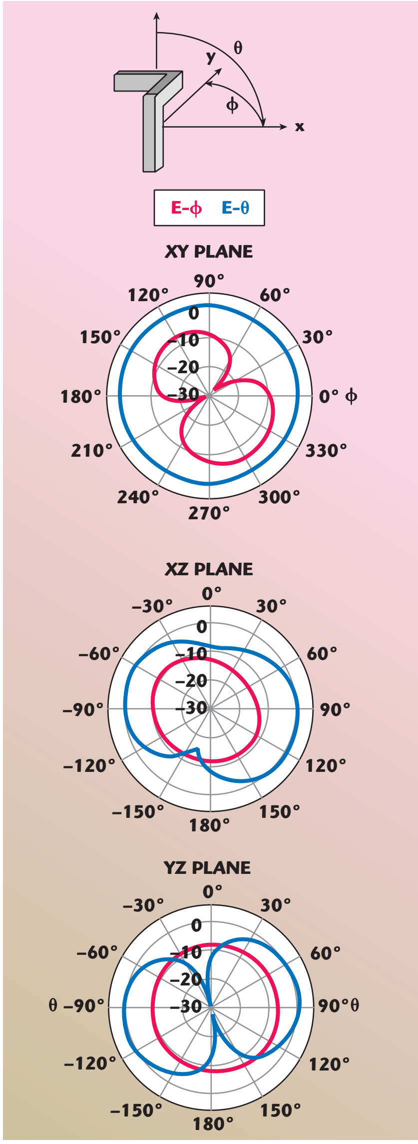


▲ *Fig. 16 Measured radiation patterns of the bent antenna.*

▲ *Fig. 17 A palm top computer with the antenna bent in three perpendicular directions.*

perpendicular directions will also reduce the effect of the human body and the environment on the antenna. Since the new antenna has a very small cross-section area and is made of a flexible material, bending the antenna in three perpendicular directions can be easily accomplished depending on the form factor of the wireless equipment. For example, the new antenna can be folded in three perpendicular directions inside notebook, laptop and palmtop computers, as shown in ***Figure 17***. It is clear that the antenna configuration can be fully embedded inside the portable computer. The calculated radiation patterns of the new antenna while it is folded in three perpendicular directions are shown in ***Figure 18***. The antenna is sensitive to all polarizations in all planes.

On the other hand, the form factor of some wireless equipment does not allow the new antenna to be internally folded in three perpendicular directions, as in cellular phones.



▲ *Fig. 18 Calculated radiation patterns of a bent antenna in three perpendicular directions.*

In such cases, special methods have to be found in order to bend the new antenna in three perpendicular directions. In one of these methods, a part of the antenna is embedded inside the handset while the other part is kept external to the handset. The internal part of the antenna is folded in two perpendicular directions parallel to the handset. The external part of the antenna is retractable and can be used as a mechanical support for the handset while it is used as a mobile TV. In this case, the external part of the antenna is mechanically supported by a thin plastic rod in order to make it mechanically rigid.

CONCLUSION

A novel wideband antenna for mo-

bile digital TV has been developed. The new antenna covers a bandwidth of more than 68 percent. It is an unbalanced resonant antenna that does not need any tuning or matching circuits. It resonates from 470 to 960 MHz and hence can cover all the UHF bands of mobile digital TV (470 to 860 MHz or a portion of it), 700 MHz WiMAX band (698 to 806 MHz), CDMA/TDMA/GSM800 (824 to 894 MHz) and E-GSM900 (880 to 960 MHz). The new antenna can be used with cellular phones, palmtop computers, notebook computers, laptop computers or any other portable communication equipment. The overall size of the new antenna is very small and its manufacturing costs are very low. Different configurations of the new mobile TV antenna have been designed, manufactured and tested. The overall sizes of the sample prototypes were 1.75 cm^3, 0.75 cm^3 and 0.5 cm^3. These were the overall sizes of the antenna because the new antenna does not require extended ground planes or any other additional components.

The antenna is made of a flexible material that can be bent and/or folded and shaped in different forms to fit the available space in any wireless equipment. It can be used as an internal, external or partially internal and partially external antenna. On the other hand, some configurations of the new antenna are aimed to be multi-polarized, which is a very important factor in MIMO and in all indoor applications. Multi-polarization can be easily achieved by bending the new antenna in three perpendicular directions. ■

Acknowledgment

The authors would like to thank Mohamed Osama for his effort in preparing this article.

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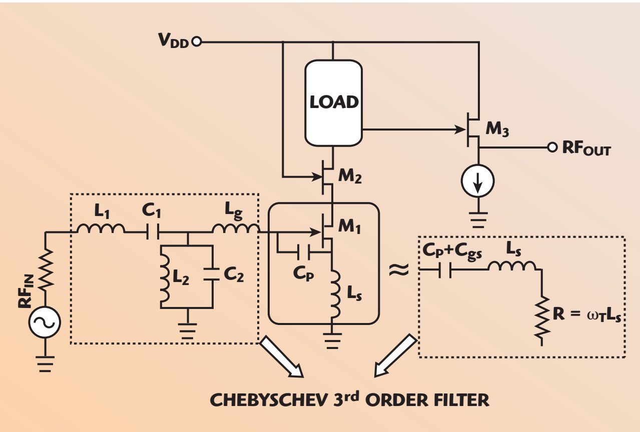
FLATNESS IMPROVEMENT FOR A SHUNT-PEAKED ULTRA-WIDEBAND LOW NOISE AMPLIFIER

A novel configuration to achieve flat gain in wideband low noise amplifiers (LNA) is presented. It consists of a conventional shunt-peaking resistor, decoupled from the cascode stage through a capacitor. A trade-off between the voltage headroom and bandwidth is obtained, improving the traditional shunt-peaking load. As an example, two LNAs for ultra-wideband (UWB), operating for mode I (3.1 to 4.8 GHz), are designed in CMOS 0.35 μm technology. The measured power gain is fairly flat, approximately 10 dB, for frequencies ranging from 3.1 to 5 GHz.

Ultra-wideband (UWB) is a technology for transmitting information spread over a large bandwidth that should be able to share spectrum with other users. This is the main reason why UWB has attracted much attention from the wireless community, both from standardization bodies and chip manufacturers. The UWB frequency range is from 3.1 to 10.6 GHz and the data rate is up to 480 Mbps.[1,2]



▲ *Fig. 1 Wideband LNA simplified schematic with wideband input impedance matching.*

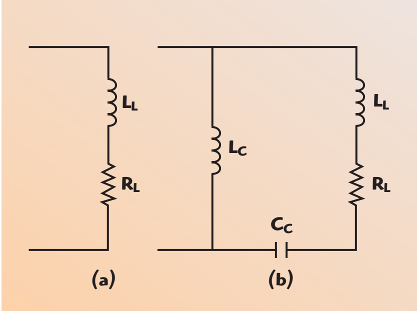
One of the most challenging components of a UWB front-end is the wideband low noise amplifier (LNA). This circuit must have a precise amplification over a wide range of frequencies. There are several methods to design a UWB LNA. Previously reported articles use the classic shunt-feedback configuration.[3-5] Another commonly used topology is the distributed amplifier,[6-8] where several gain stages are connected so that their capacitances are separated. Series inductive elements are used to separate capacitances at the inputs and outputs of adjacent gain stages occupying a high area. Finally, inductor degenerated amplifiers with LC-ladder input matching networks are also employed.[9-11] This configuration expands the use of the typical inductively degenerated narrow band amplifier, by embedding the input network of the amplifying device in a multi-

H. GARCÍA-VÁZQUEZ, SUNIL L. KHEMCHANDANI, J. ARIAS-PÉREZ AND J. DEL PINO
University of Las Palmas de Gran Canaria Spain

section reactive network. In this way, the overall input reactance is resonated over a wider bandwidth, achieving at the same time a wideband input match and low NF. To extend the gain bandwidth, the load is generally composed of a shunt-peaking resistor.[12] This technique imposes an upper limit to maximum gain and flatness. To overcome this issue, a modification of the conventional shunt-peaking load is proposed.

CIRCUIT DESCRIPTION

A schematic of the wideband input matched CMOS LNA is shown in ***Figure 1***. It consists of a wideband input matching circuit, a gain stage with inductive degeneration (L_g and L_S)[11] and a wideband output load. In order to buffer the output to an external 50 Ω load, an emitter follower (M_3) has been included. As the figure shows, the input matching circuit consists of a filter embedded with the input impedance of M_1. In this case, a third-order bandpass Chebyshev filter in T configuration was selected. In order to increase the flexibility of the filter, C_P is introduced between the gate and source of M_1.



▲ *Fig. 2 Conventional shunt-peaking load (a) and modified shunt-peaking load (b).*

The gain stage is composed of a cascode transistor, which improves the reverse isolation and lowers the Miller multiplied capacitance. The width and polarization current of the transistors are optimized for noise and power consumption. As shown in ***Figure 2***, and in order to provide a wideband operation, the typical resonant load used in narrow band LNAs is replaced by a shunt-peaking resistor.[12] With this configuration the overall amplifier gain should be flat across the pass-band. The amplifier gain is given by the product of the transistor transconductance (gm) and the magnitude of the impedance of the shunt-peaking load, given by:

$$Z_L(j\omega) = \frac{R_L + j\omega L_L}{1 - \omega^2 L_L C_{out} + j\omega C_{out} R_L} \quad (1)$$

where C_{out} represents the equivalent capacitance at the output node, which includes the transistor output capacitance, the loading by interconnections and subsequent stages, and the parasitic capacitance of the inductor. This expression contains a zero and two complex poles. The extended bandwidth comes from the $|Z(j\omega)|$ increase due to the poles below the $L_L C_{out}$ resonance ($\omega_0 = 1/L_L C_{out}$) and to the zero ($\omega_z = R_L/L_L$). Unfortunately, this leads to a peak in the frequency response, thus degrading the flatness. A possible solution is to keep both resonances out-of-band by using a low value of L_L, which in turn implies a low gain.

In order to have a large gain, R_L should be chosen sufficiently high to improve the gain at lower frequencies. However, the voltage headroom imposes an upper limit to R_L and, as a consequence, to maximum gain and flatness. To overcome this issue, a modification of the conventional shunt-peaking load is proposed.

The proposed shunt-peaking load is also shown in the Figure. It is based on a conventional shunt-peaking resistor, decoupled from the cascode stage through a capacitor C_C. To bias the active stage, an inductance L_C is placed between V_{DD} and the M_2 drain. The impedance of the new shunt-peaking load is given by

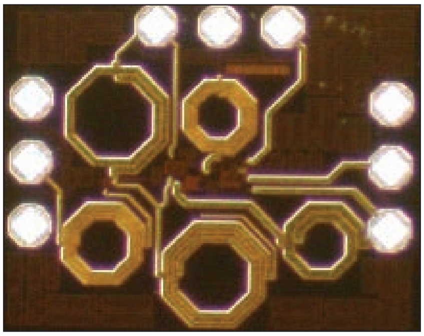
$$Z(j\omega) = \frac{j\omega L_c R_L\left(\dfrac{j\omega L_L}{R_L}+1\right)}{1 - j\omega^3 L_c L_L C_{out} - \omega^2 L_c R_L C_{out} + j\omega(L_c + L_L)} \quad (2)$$

The C_C value has been chosen high; consequently, its effect is neglected and it does not appear in Equation 2. With this configuration, R_L can be chosen higher than in a conventional shunt-peaking load, overcoming the voltage headroom limitation. The immediate consequence is that a flatness improvement is achieved.

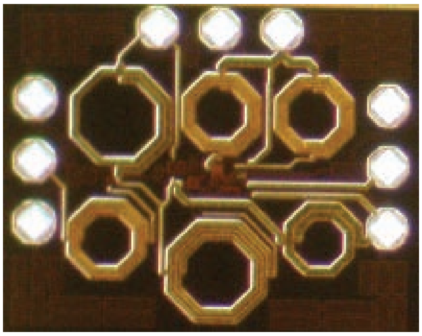
DESIGN EXAMPLE AND COMPARISON

To demonstrate the practical viability of the proposed structure in CMOS technology, both the proposed topology and the conventional one have been applied to a 3.1 to 4.8 GHz wideband amplifier, based on a 0.35 μm standard CMOS process. Both circuits were optimized with the pads and on-chip spiral inductors analyzed with the MOMENTUM electromagnetic simulator.[13]

Figures 3 and ***4*** show the photos of the LNA with shunt-peaking load and the LNA with modified shunt-peaking load, respectively. As can be observed, the layouts are very similar with the exception of the L_C inductor in the upper right corner. In both designs, the chip size, including the probe pads, is 949 × 760 μm. Each amplifier draws 17 mA from a 3.3 V supply.



▲ *Fig. 3 Photograph of the LNA with shunt-peaking load.*



▲ *Fig. 4 Photograph of the LNA with the modified shunt-peaking load.*

The measured forward gain and input return loss of the amplifiers are shown in ***Figure 5*** for frequencies from 2 to 6 GHz. For the proposed shunt-peaked LNA, the power gain is fairly flat at

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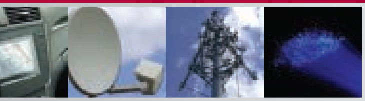

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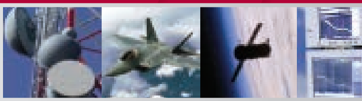

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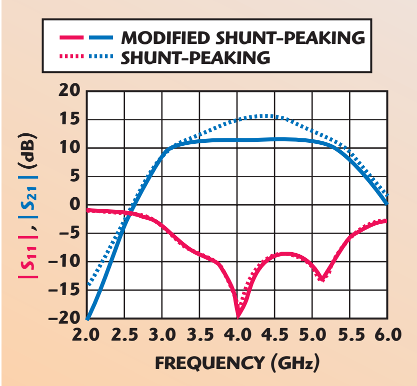
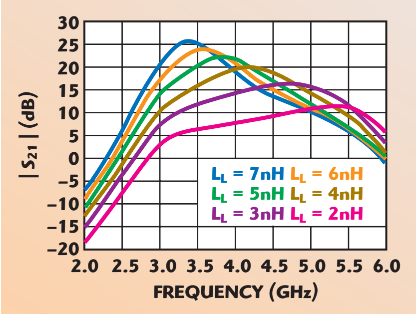
approximately 10 dB for frequencies ranging from 3.1 to 5 GHz. However, the same cannot be said for the conventional case, where a peak is evident. To flatten the insertion gain, the zero pole frequency should be placed as close as possible to the upper edge of the band by lowering L_L. However, this entails a gain reduction, as can be seen in ***Figure 6***, where the simulated S_{21} is plotted for different ideal L_L inductors. As the inductance decreases, the flatness and bandwidth increase but the gain drops.

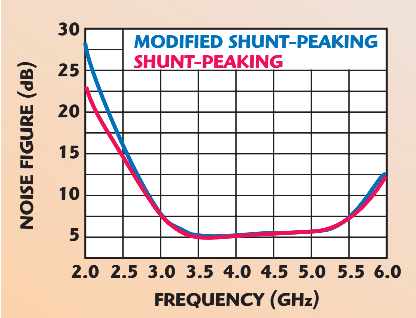
In both amplifiers, the input return losses remain the same, because both circuits share identical input matching circuits. As a consequence, the noise figure (see ***Figure 7***) is also the same in both cases.



▲ *Fig. 5 Measured S-parameters for LNA with shunt-peaking load and modified shunt-peaking load.*



▲ *Fig. 6 Gain simulation for different L_L inductance using conventional shunt-peaking.*



▲ *Fig. 7 Measured noise figure for LNA with modified shunt-peaking and shunt-peaking load.*

CONCLUSION

In this article, a novel wideband load configuration is presented. It is based on a conventional shunt-peaking resistor decoupled from the cascode stage through a capacitor. It offers the advantages of broadband operation and substantially higher gain flatness than a conventional shunt-peaking load. Its practical implementation in a standard low cost 0.35 µm process was discussed and verified. The presented results make the LNA design suitable for the 3.1 to 4.8 GHz UWB frequency range. ■

ACKNOWLEDGMENT

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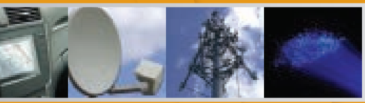

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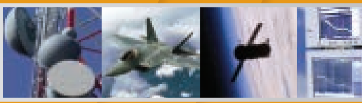

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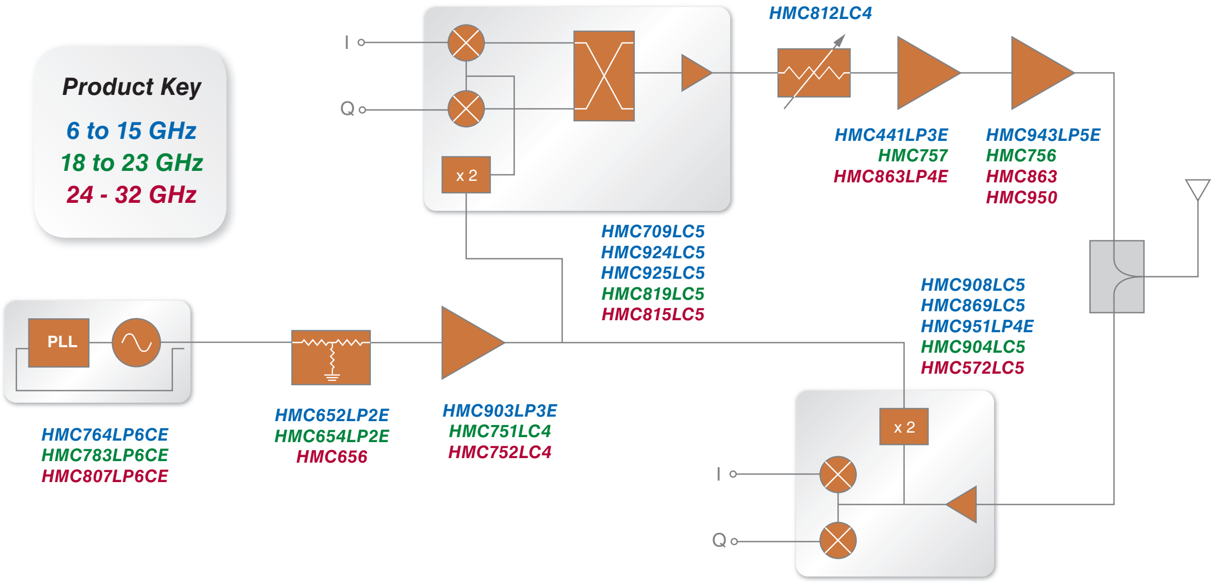

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A Compact, Balanced Low Noise Amplifier for WiMAX Base Station Applications

This article presents the design and development of a balanced low noise amplifier suitable for WiMAX base station applications. The balanced low noise amplifier employs a compact 3 dB branch line coupler and advanced GaAs E-PHEMT technology to realize excellent return loss, noise figure and linearity performance as well as compact size. By using the compact branch line coupler, a size reduction of 40 percent is achieved as compared to using the conventional branch line coupler. The balanced LNA, using a compact branch line coupler designed on a printed circuit board PCB with Rogers 4003C substrate, exhibits a small-signal gain of 14.30 dB at 2.3 GHz and 13.50 dB at 2.7 GHz, a noise figure less than 1 dB, an OIP3 of 37.2 dBm and a P1dB of 21.28 dBm at 2.5 GHz.

Base station applications for WiMAX require low noise amplifiers (LNA) with low noise figures, high linearity, compact size, and excellent input and output return loss.[1] A balanced amplifier topology is often the preferred choice in LNA applications for base station transceiver front-ends, because it presents several advantages over single-ended amplifiers such as excellent input and output return loss, higher linearity and better stability.[2] These requirements present several design challenges that involve technology and design trade-offs of couplers to achieve an optimal solution in terms of size and performance. GaAs E-PHEMT has an enormous potential for realizing low noise amplifiers at microwave frequencies, due to their low noise features, high electron velocity and positive gate bias voltage requirements.[3]

A branch line coupler is one of the most popular hybrids for the design and implementation of balanced LNAs.[4] However, a branch line coupler requires a large circuit area, due to the quarter wavelength transmission lines.[4] Various efforts to reduce the size of the branch line coupler have been reported.[5-7] The Lange coupler[8] is relatively impractical for fabrication on a printed circuit board (PCB) as the lines are very narrow, close together and is more practical for MMIC designs. The broadband lumped element 3 dB quadrature coupler[9] is more suitable for MMIC applications. A branch line coupler is more suitable for base station LNAs because it has better input and output return losses and lower insertion loss, compared to the lumped element approach. In order to design a reduced

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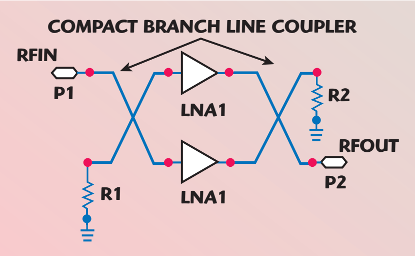
size branch line coupler on a PCB for low cost applications, the design technique proposed by Ching-Wen Tang[10] is more promising and hence chosen for this design.



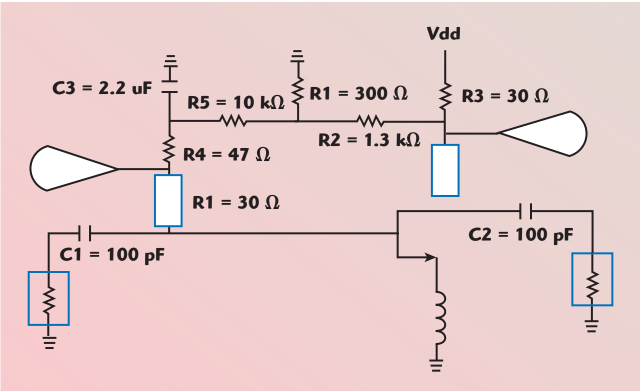
▲ *Fig. 1 Simplified schematic of the proposed balanced LNA.*

BALANCED LNA DESIGN

The schematic of the proposed balanced LNA is shown in ***Figure 1***. The design of a balanced LNA comprises two main steps. The first step involves the design of a single stage LNA. Once the single stage amplifier has been designed, the next step involves the design of an appropriate coupler to meet the size and bandwidth requirements.

LNA DESIGN

The Avago Technologies ATF-54143[11] high dynamic range, low noise enhancement mode PHEMT (E-PHEMT) device was selected for the amplifier design. The schematic of the LNA is shown in ***Figure 2***. A voltage divider, consisting of R1 and R2, is used to bias the FET. The purpose of R4 is to enhance the low frequency stability of the device by providing a resistive termination at low frequencies. The capacitor C3 provides a low frequency bypass for R4. Additional resistance in the form of R5 (approximately 10 kΩ) is added to provide current limiting for the gate of enhancement mode devices such as the ATF-54143. Quarter wavelength microstrip lines are used to provide the DC feed and to improve the capability of blocking the RF signal.



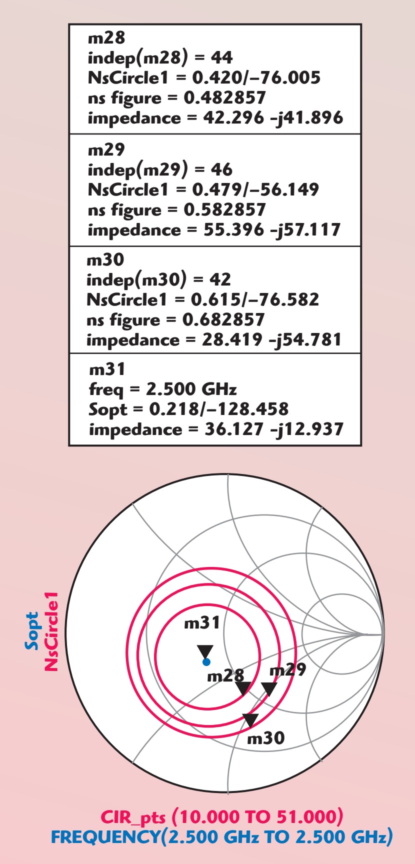
▲ *Fig. 2 Schematic of the proposed LNA.*

Radial stubs are used, together with the quarter wavelength microstrip lines, to improve the isolation of the DC and RF path.[4] The LNA stability is improved by adding an inductance at the emitter (source) of the active device.[12] Input impedance matching, using inductive source degeneration, is a popular approach, because matching to the source does not introduce additional noise (as in the case of using a shunt input resistor) and does not restrict the value of g_m (as in the case of the common-gate configuration).

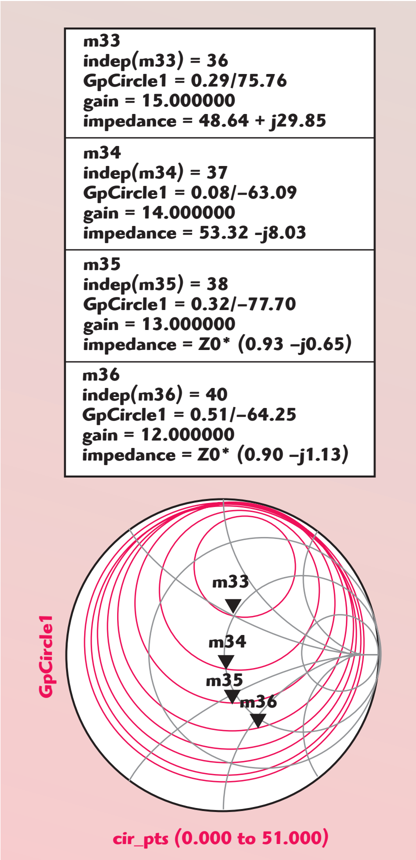
INPUT AND OUTPUT MATCHING

For low noise applications, the input matching is done at Γ_{opt} to achieve the lowest noise possible.[4] The maximum gain is achieved, for unconditionally stable amplifiers, when both input and output are conjugate matched. The results of Γ_{opt} and noise circles, calculated for the input matching design, are shown in ***Figure 3***.

For output matching requirements, the G_P circles calculated for various gain values are shown in ***Figure 4***. In order to achieve the lowest LNA noise figure possible, the input stage of the amplifier is matched to Γ_{opt},[4] the location of which is 36.127-j12.93 Ω. The matching network is designed to match 50 Ω to 36.127- j12.93 Ω. A single stub matching technique, using series and shunt open stubs, is selected to design the input matching network. The output matching network transforms the load impedance from 48.64 + j29.86 Ω to 50 Ω for a gain of 15 dB.



▲ *Fig. 3 Simulated Γ_{opt} and noise circles.*



▲ *Fig. 4 Simulated G_p circles.*

COMPACT BRANCH LINE COUPLER DESIGN

The size reduction of the branch line coupler is practical, using the approach proposed by Razavi.[12] ***Figure 5*** shows the equivalent circuit of the quarter wavelength transmission line. Three types of shortened quarter wavelength transmission line can be used for the compact branch

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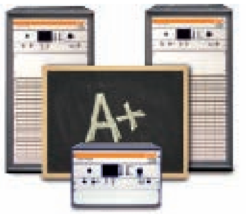

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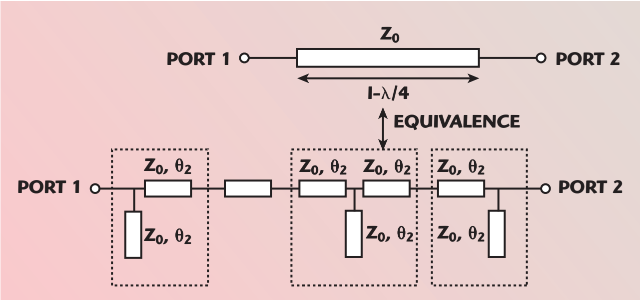
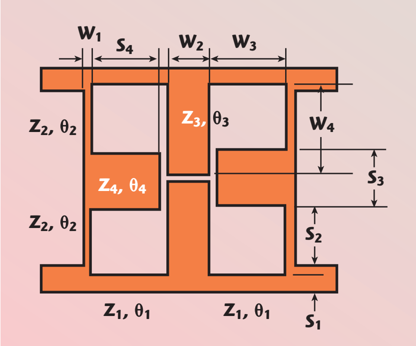
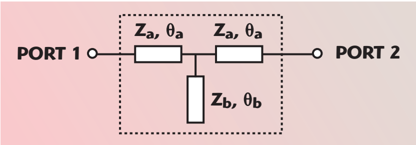


▲ *Fig. 5 Equivalent circuit of a quarter wavelength transmission line.*



▲ *Fig. 6 Diagram of the T model microstrip branch line coupler using a low impedance approach.*



▲ *Fig. 7 Equivalent quarter wavelength transmission line of T model.*

line couplers. They are the T-model, the π-model and a combination of both models.[12] The low impedance T-model approach is selected in this research, due to the ease of implementation and the reduced complexity in terms of layout. The design uses the low impedance open stub in the shortened quarter wavelength transmission line of the T-model. ***Figure 6*** shows the T-model microstrip branch line coupler using the low impedance approach.

$$Z_a = \frac{Z_0}{\tan\theta_a} \quad (1)$$

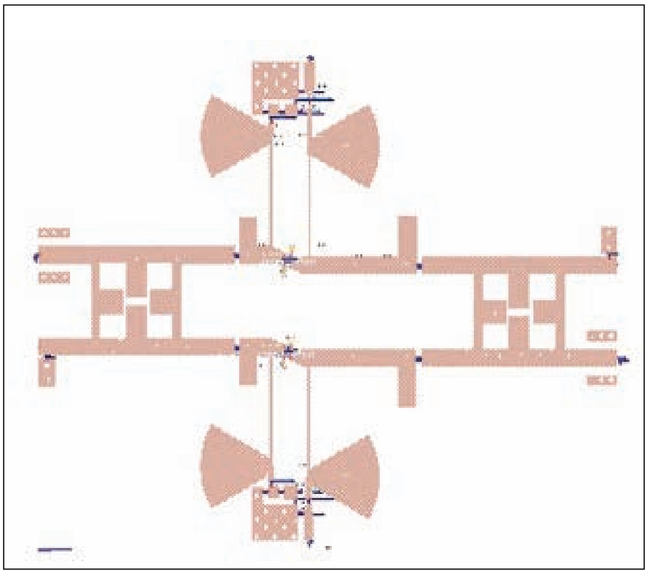
$$Y_b \tan\theta_b = \frac{2}{Z_a \tan 2\theta_a} \quad (2)$$

Using Equations 1 and 2, the impedance and lengths of each shortened equivalent quarter wavelength, which are Z_a, Z_b, θ_a and θ_b as depicted in ***Figure 7***, are derived for both the shunt and series arms. A size reduction of approximately 40 percent is reported using a compact branch line coupler. The complete design of the balanced LNA, using compact branch line couplers, is shown in ***Figure 8***.

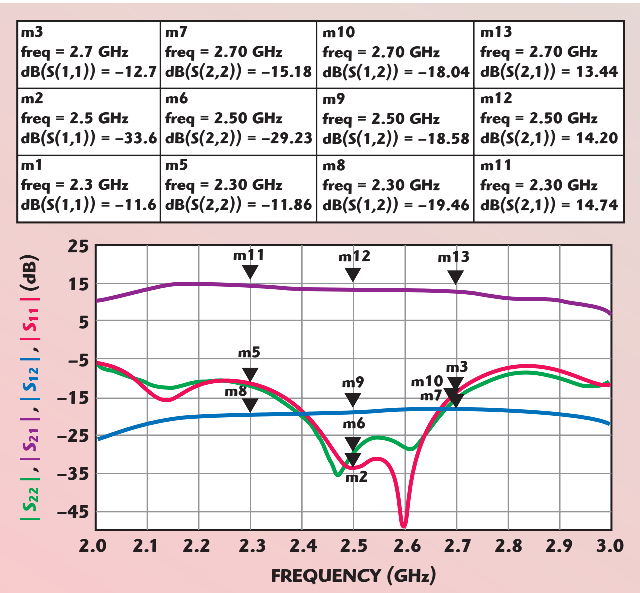
SIMULATION AND MEASURED RESULTS

The simulated small-signal S-parameters, which includes the gain, input and output return losses and isolation, are shown in ***Figure 9***. As can be observed, the maximum value of gain obtained is 14.74 dB at 2.3 GHz and rolls off gradually to 13.44 dB at 2.7 GHz. The input and output return losses are very good across a limited bandwidth, better than 10 dB from 2.3 to 2.7 GHz. The measured small-signal S-parameters are shown in ***Figure 10***. Good agreement between the simulated and measured results can be observed. The measured gain is slightly lower, 14.30 dB at 2.3 GHz to 13.50 dB at 2.7 GHz. Both simulation and measured results show a gain flatness of less than 1 dB. The input and output return losses degraded slightly, compared to the simulated results. The reason for the deviation of a few dB is due to the losses in the microstrip elements of the amplifier and coupler circuits, variations in the SMD devices and coupling of signals between the amplifiers. The comparison between simulated and measured noise figure results is shown in ***Figure 11***.

The noise figure measurements were made in a shielded room. The measured noise figure performance was found to degrade slightly, compared to the simulated noise figure of the circuit. However, the measured noise figure is well below 1 dB and only degrades approximately 0.2 to 0.3 dB from simulation. This low noise figure has been achieved by matching the input at Gamma Opt and using E-PHEMT device. The simulated noise figure is typically lower than the measured one, due to unaccounted sourc-



▲ *Fig. 8 Layout of the balanced LNA using compact branch line couplers.*



▲ *Fig. 9 Simulated S-parameter results.*


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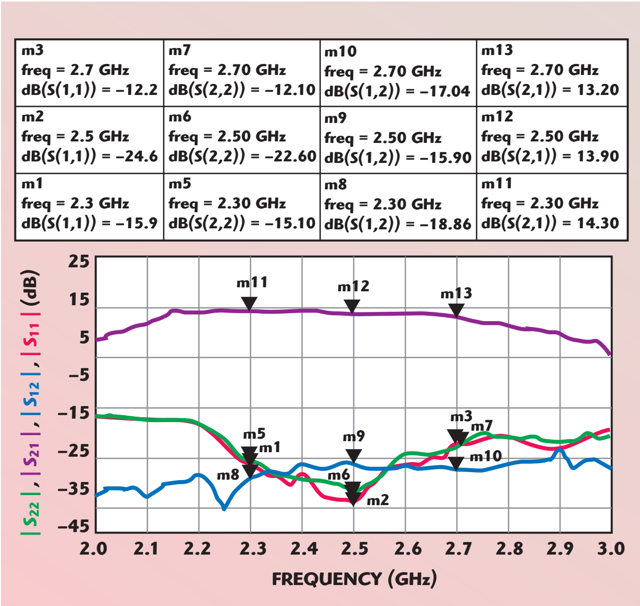
TABLE I

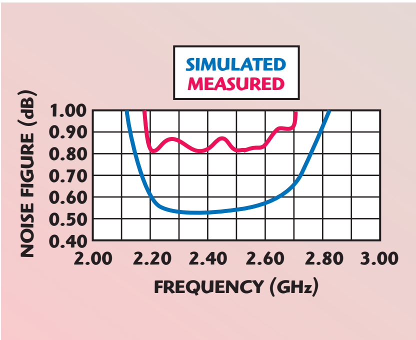
COMPARISON BETWEEN MEASURED AND SIMULATED RESULTS AND OTHER REPORTED LNAs

References	Frequency (GHz)	Gain (dB)	NF (dB)	IIP3 (dBm)	P1dB (dBm)	Pdis (mW)
This Work (Measured)	2.3-2.7	13.20	0.92	23.80	21.28	192.7
This Work (Simulated)	2.3-2.7	13.44	0.65	21.70	21.61	185
(13)	2.44	15.3	3.34	-10	-18	3.9
(14)	2.4-6	13.0	2.5	-3.9	-	6
(15)	0.8-1.7	37	0.9	-	-	-
(16)	0.1-8	16	3.4-5.8	-9	-	16
(17)	2-10	10	4.6	-	-	100
(18)	2.5-4	10.6	4	-8	-	8

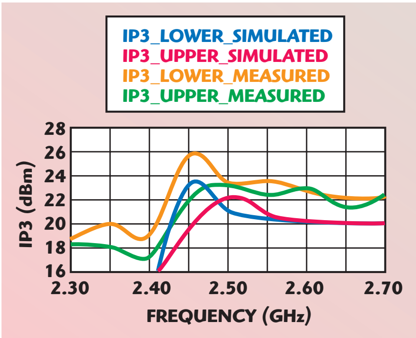
es of noise such as the noise contributed by the power supply and ambient noise. ***Figure 12*** shows the simulated versus measured input IP3. The higher input IP3 is achieved using the balanced approach as compared to the single-ended design. Measurement and simulated results are summarized and compared with previous reports of LNA designs in ***Table 1***. The complete fabricated device is shown ***Figure 13***.



▲ *Fig. 10 Measured S-parameter results.*



▲ *Fig. 11 Simulated and measured noise figure.*



▲ *Fig. 12 Simulated and measured input IP3 results.*

CONCLUSION

A balanced LNA design, which uses compact branch line couplers, has been designed and developed. This LNA exhibits a small-signal gain of 14.30 dB at 2.3 GHz and 13.50 dB at 2.7 GHz. An output power of 21.28 dBm and an OIP3 of 37.2 dBm at 2.5 GHz have been obtained. Excellent noise figure performance below

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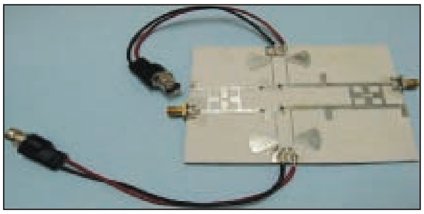

▲ *Fig. 13 Photograph of the fabricated LNA.*

1 dB is achieved across the bandwidth of interest. This design is suitable for applications that require a smaller size device. By using the compact branch line couplers, a size reduction of 40 percent is achieved as compared to using the conventional branch line coupler. Excellent performance could be achieved even by reducing the branch line coupler by almost 40 percent. ■

ACKNOWLEDGMENT

The author would like to thank the Universiti Kebangsaan Malaysia and USM for their support.

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DIAMOND RF™ RESISTIVES: THE ANSWER TO HIGH POWER AND LOW CAPACITANCE

The technology in this article is used in the high power Diamond Rf™ Resistives—resistors, terminations and attenuators—from EMC Technology. These components demonstrate excellent electrical and thermal performance due to the high thermal conductivity of the CVD diamond substrate material, which results in a component size reduction for a given power dissipation. Components built with this material also demonstrate lower parasitic behavior (i.e. better isolation), which is ideal for applications such as phased array radar. The characteristics of the CVD diamond material impacts performance and design for applications such as high power Wilkinson power dividers/combiners, dual junction circulator duplexers and feedback networks for power amplifiers.

A single 0402 resistor, manufactured on a CVD diamond substrate, can dissipate 20 W of continuous electrical energy and 200 W of pulsed energy. This provides great flexibility to the design engineer, who is continually asked to reduce size and increase performance. This article describes the current power resistor implementations, the limitations of conventional power resistors and shows how diamond passive components can provide a path to reduce the limitations in real applications.

Properly dissipating heat in resistive RF and microwave components has always been a challenge for design engineers. A thin film resistor, required to dissipate large amounts of heat, needs a substrate with high thermal conductivity to transfer the heat to the system heat sink. Over the past few decades beryllium oxide (BeO) and aluminum nitride (AlN) have been the preferred substrate materials for high power RF resistors. These ceramic materials have relatively high thermal conductivities and suffice for many applications; however, as designers face size reduction challenges, the BeO and AlN capabilities become inadequate. This leaves designers seeking alternative methods to dissipate heat in a small footprint. Type 2 CVD diamond has been developed as a substrate material to address the shortcomings of BeO and AlN; a material with a thermal conductivity eight times greater than AlN.

CVD diamond is a single crystal carbon substrate material that is produced using chemical vapor deposition (CVD) processing. From a structural view, Type 2 CVD diamond is similar to the diamond that can be found naturally in the earth. Due to the nature of the material, it can be processed similarly to other substrate materials to form electrical circuits. Thin film resistors can be developed by sputtering tantalum nitride (TaN) on to the CVD diamond substrate. Conductor materials can also be sputtered onto CVD diamond in order to create eutectic solder pads or wire bond pads allow-

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TABLE I

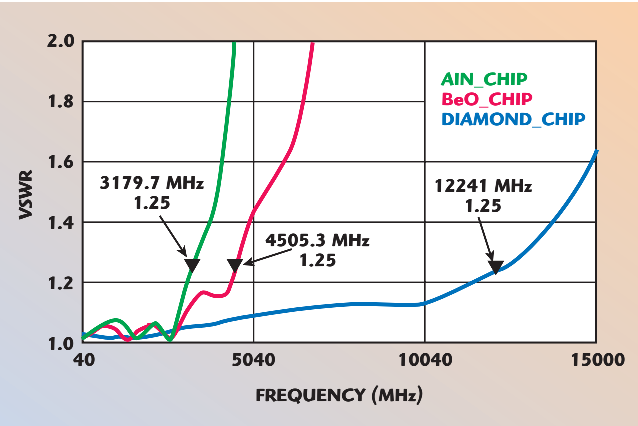
THERMAL AND RF PROPERTIES OF SUBSTRATE MATERIALS

Material	Thermal Conductivity (W/cm°C)	Dielectric Constant	Loss Tangent
AlN	1.7	8.9	0.0005
BeO	2.6	6.5	0.0040
CVD Diamond	16.0	5.6	0.0005

TABLE II

THIN FILM RESISTOR DIMENSIONS

Device	Substrate Material	Substrate Thickness (inches)	Length (inches)	Width (inches)	Area ($inches^2$)	Cap./W (pF/W)
CT1310D	CVD Diamond	0.015	0.068	0.068	0.005	0.003
82-3031	BeO	0.040	0.160	0.130	0.021	0.007
82-7176	AlN	0.040	0.230	0.120	0.028	0.012



▲ *Fig. 1 VSWR vs. frequency for different terminations.*

ing CVD diamond circuits to be easily used in many mounting configurations, such as surface mount, chip and wire, flange mount, etc.

While CVD diamond is well known for its durable mechanical strength, its thermal conductivity is where it shines for RF resistive components. ***Table 1*** compares the thermal conductivities and RF properties of BeO, AlN and CVD diamond. The excellent RF and thermal properties of CVD diamond make it a great choice for thin film resistive products for demanding, high reliability applications such as military, space and wireless infrastructure.

FIGURE OF MERIT: CAPACITANCE PER WATT

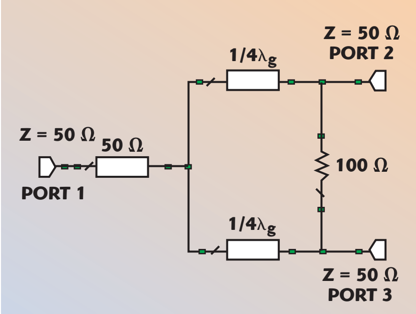
Since CVD diamond has a thermal conductivity that is approximately eight times higher than AlN, a CVD diamond-based thin film resistor will need eight times less area to dissipate the same amount of heat. The physical size of a power resistor is directly related to the thermal conductivity of the substrate. A low thermally conductive substrate requires more resistor surface area. The increased size of the device presents two problems for microwave designers:

1. More PCB space is required
2. The upper frequency is limited due to the increased capacitance

The increased capacitance is simply due to the parallel plate capacitance between the thin film material and the ground plane under the chip. A helpful figure of merit is the capacitance per watt (C/W). For two resistors with the same power rating, a lower C/W will indicate that the resistor's upper frequency range will be less constrained by capacitance. This is becoming increasingly important for systems requiring wideband frequency responses, as the reactive tuning needed for diamond is eight times less difficult than for traditional AlN resistors.

To highlight the significance of lower C/W, here is an example of three solutions to handle 120 W of dissipated power:

1. CT1310D (CVD diamond 50 Ω termination)
2. 82-3031 (BeO 50 Ω termination)
3. 82-7176 (AlN 50 Ω termination)



▲ *Fig. 2 Traditional Wilkinson divider/combiner.*

Table 2 compares the physical sizes of each chip thin film resistor. As seen from the Table, the C/W ratio of the CVD diamond chip is significantly lower than that of conventional substrate materials.

The reduced capacitance per watt of CVD diamond has many advantages for RF circuits and also translates to higher frequency of operation. ***Figure 1*** shows the measured VSWR of these three chip terminations. For comparison purposes, the frequency at which the VSWR increases above 1.25 is noted. The CVD diamond operates up to 12.4 GHz before crossing the VSWR limit. The AlN chip operates to 3179 MHz before crossing the VSWR limit. The BeO chip operates to 4505 MHz before crossing the VSWR limit.

COMMON APPLICATIONS OF DIAMOND RF RESISTIVES

Due to the distinct benefits of CVD diamond, it is worthwhile to present some common applications where this product line has solved design engineering problems, specifically Wilkinson combiners/dividers, isolator circuits and amplifier feedback networks.

High Power Wilkinson Power Dividers/Combiners

Figure 2 shows the highly popular single stage Wilkinson divider/combiner.[1] High power dividers/combiners require a high power resistor to provide isolation between the combining ports. The isolation resistor can see large amounts of power if the signal at each leg (ports 2 and 3) are out of phase or have unequal amplitudes. The worst case occurs when the signals are 180° out of phase, in which case the isolation

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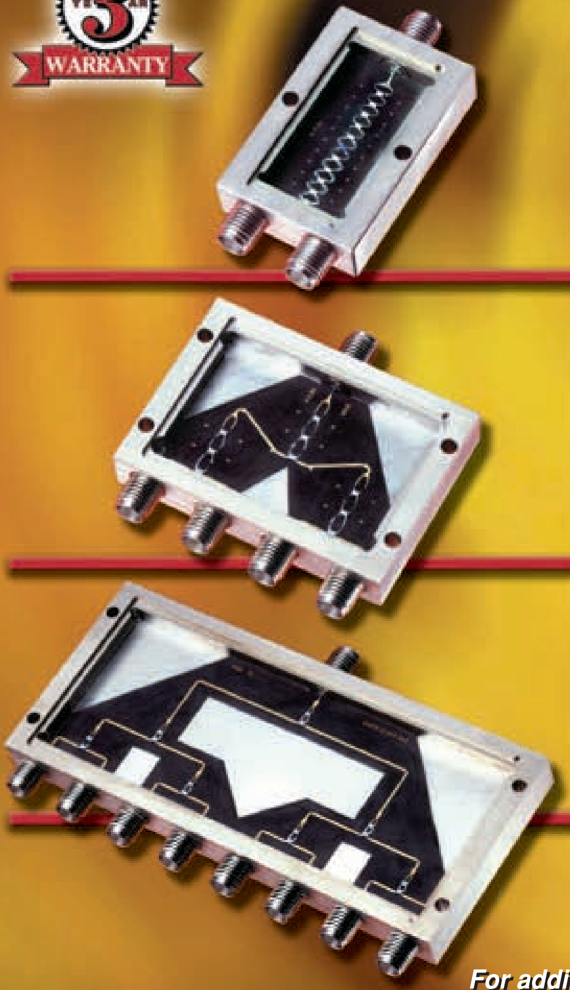

RF INPUT PARAMETERS	UNITS	MINIMUM	MAXIMUM
2 Way Power Divider - Model PD02-18004000			
RF frequency range	GHz	18	40
Insertion loss	dB		1.5
Isolation	dB	17	
Input VSWR	Ratio		1.8
Output VSWR	Ratio		1.7
Phase unbalance	Degrees		±5.0
Amplitude balance	dB		±0.5
4 Way Power Divider - Model PD04-18004000			
RF frequency range	GHz	18	40
Insertion loss	dB		2.5
Isolation	dB	17	
Input VSWR	Ratio		1.8
Output VSWR	Ratio		1.7
Phase unbalance	Degrees		±5.0
Amplitude balance	dB		±0.5
8 Way Power Divider - Model PD08-18004000			
RF frequency range	GHz	18	40
Insertion loss	dB		3.5
Isolation	dB	17	
Input VSWR	Ratio		1.8
Output VSWR	Ratio		1.7
Phase unbalance	Degrees		±5.0
Amplitude balance	dB		±0.5

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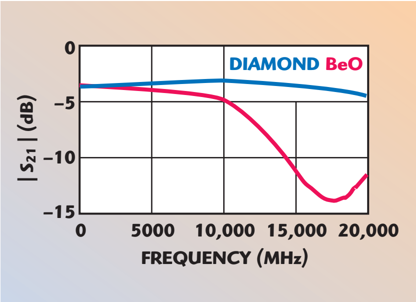
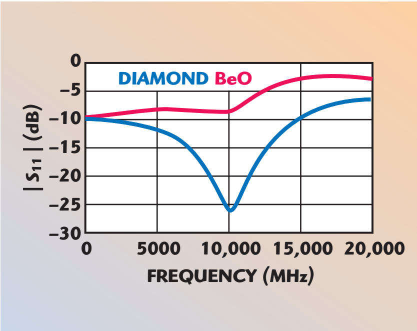
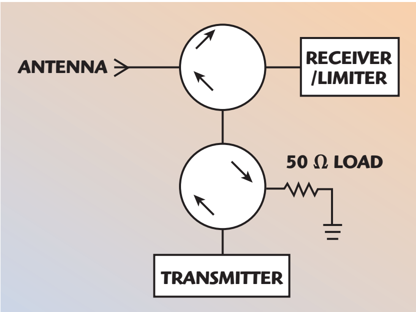


▲ *Fig. 3 S_{21} vs. frequency for a Wilkinson combiner.*



▲ *Fig. 4 S_{11} vs. frequency for a Wilkinson combiner.*



▲ *Fig. 5 Dual junction circulator duplexer.*

resistor will dissipate a majority of the energy. As the trend continues to make RF circuits smaller, the resistor hits a limitation of power dissipation and upper frequency operation. The size reduction of the traditional BeO or AlN resistor is limited due to the substrate thermal conductivity. For a given power rating, the CVD diamond resistor can be made significantly smaller. As shown in Table 2, a 150 W CVD diamond part will have a capacitance per watts ratio of less than half that of the BeO part. This reduced parasitic capacitance allows the combiner with the CVD diamond isolation resistor to operate at a higher frequency with significantly less loss and greater isolation.

For comparison purposes, two Wilkinson combiners were designed for a center frequency of 10 GHz (X-band). One combiner uses a 50 W BeO 100 Ω resistor and the other uses a 50 W CVD diamond 100 Ω resistor. ***Figures 3*** and ***4*** show the S_{21} and S_{11} results, respectively. As expected, the BeO resistor has a degraded performance at higher frequencies, due to the shunt capacitance introduced by the resistive film physical size. If using the BeO isolation resistor, a reactive tuning is required to tune out the capacitance for optimized insertion loss and isolation, resulting in a very narrow band combiner and manufacturing repeatability challenges. Diamond resistors have a distinct advantage in this application.

Dual Junction Circulator Duplexer Applications

Typical phased array radars, depending on the number of elements, can have hundreds or even thousands of transceivers. Many transceiver topologies require a duplexer for switching between transmit and receive modes. In many cases, the duplexing function is implemented with an isolator in order to isolate the sensitive receive side from any high power transmit crosstalk. In some situations, a dual junction circulator is used to provide isolation up to 50 dB (see ***Figure 5***).[2]

With this topology, any power that leaks from the transmitter will be absorbed by the 50 Ω termination. The termination resistor must be capable of handling the maximum transmit power, minus the isolation of the isolator. These peak power levels can be extremely high in many radar applications. In the past, these resistors would be large BeO or AlN resistors. The CVD diamond resistor can be used in place of the ceramic resistors. With a smaller size diamond resistor and inherently less capacitance, the duplexer module can be made smaller and the isolator termination port can be better matched to prevent damage to the transmitter. The decreased size of the duplexer is multiplied by the number of array elements, producing a meaningful system size reduction.

Additionally, the decreased capacitance of CVD diamond terminations creates an ideal solution for high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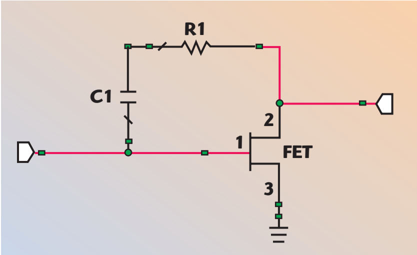

▲ *Fig. 6 Typical FET power amplifier with negative feedback.*

frequency radar. As expected, the VSWR of BeO and AlN terminations is degraded at higher frequencies and can ultimately limit the overall performance of the isolator.

In some instances, the 50 Ω termination used for the isolation load can be replaced with a 20 or 30 dB attenuator. This technique is used for power sampling of the power dissipated by the attenuator. This power level can be used as a method of measuring power incident to the attenuator. Similar to the load resistor, the attenuator needs to be sized to handle the worst case reflected power. If implemented with CVD diamond, a size reduction of up to 14 times can be achieved.

Power Amplifier Feedback Resistor

As military and commercial wireless applications begin to stretch across multiple bands, power amplifiers have been required to operate over wider bandwidths. As the bandwidth requirements increase, negative feedback stability compensation becomes necessary.[3] ***Figure 6*** shows a simple FET power amplifier with negative feedback. The feedback provides a method of bandwidth improvement and a method of stabilization. In a previous article, the author describes a 0.7 to 1.8 GHz amplifier, which uses a 10 W GaN HEMT and negative feedback.[4] The design requires that the feedback resistance be split into two 1206 size resistors, in order to enhance the power handling capabilities. While this may suffice for PA applications to 10 W, higher power amplifiers will require significantly more dissipation capabilities from the feedback resistors. In another article, the author describes a 100 W amplifier using LDMOS transistors with a flanged feedback resistor.[5] In this case, the large feedback resistor requires a flanged device to handle the amount of feedback power dissipated in the resistor. This can lead to a physically large design and an increase in shunt capacitance of the resistor, which will load the output of the LDMOS amplifier.

In both amplifier designs, the CVD diamond resistors can offer benefits of greater power handling in a smaller package than the ceramic counterparts resulting in an overall size reduction.

CONCLUSION

The key system parameters of radar and wireless infrastructure systems are pushing the capabilities of even the simplest of passive components, the resistor. These passive components need to handle more power, at higher frequencies, and with smaller packages. The Diamond Rf™ Resistive technology is a proven solution to equip wireless designers with a resistive component that handles more power, at higher frequencies, and is produced in a smaller package. ■

For more information:
http://www.emc-rflabs.com/Passive-Components/Overview.aspx

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VALIDATING A 3 GHZ CALIBRATION KIT FOR TESTING TO 6 GHZ

With the introduction of high-definition video signals operating at data rates up to 1.485 Gb/s for 1080i and 2.97 Gb/s for 1080p, a new generation of BNC connectors needs to perform at significantly higher frequency than standard BNCs in broadcast studio and transmission applications. To meet the broadcast industry's stringent HDTV standards such as SMPTE 292M and 424M, Radiall has designed a new 75 ohm HDTV BNC connector that offers a true 75 ohm interface with low VSWR and return loss over a frequency range from 0 to 6 GHz.

The new connector design needs to keep a suitably low VSWR to ensure signal integrity. The increased frequency range—from 3 to 6 GHz—presents concerns over testing the connectors. Commercially available 75 ohm calibration kits are specified only to 3 GHz. However, kits should be usable at higher frequencies once they are validated to such frequencies. This article explains how Radiall validated its existing 75 ohm calibration kit for performance to 6 GHz. Specifically, Radiall wanted confidence that it could accurately measure VSWR and return loss for the connector.

Calibration kits are essential to adjusting a vector network analyzer (VNA) and other test equipment for accurate measurements. A kit contains precision components of known characteristics to allow the VNA to be accurately calibrated and to remove the effects of adapters and other components from the results. Once calibrated, the VNA reveals valuable information about a component's VSWR, return loss, impedance and other characteristics by evaluating reflected and transmitted energy.

To validate the 75 ohm calibration kit at higher frequencies, Radiall compared its performance to that of a 50 ohm kit at frequencies

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up to 6 GHz. Since the 50 ohm kit is specified up to 18 GHz, comparing the VSWR of the two kits allows determination of the suitability of using the 75 ohm kit to characterize the new connectors at high frequencies.

50 AND 75 OHM CALIBRATION DYNAMIC MEASUREMENTS

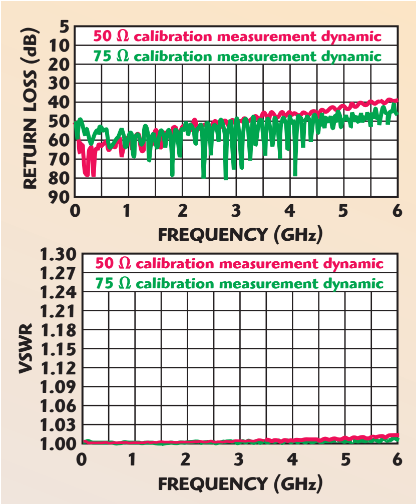
First, the RF dynamics of the two kits were evaluated by measuring return loss and VSWR. As shown in ***Figure 1***, the results were quite close. For broadcast market testing needs, a dynamic lower than -50 dB between 0 to 3 GHz and -40 dB between 3 to 6 GHz are acceptable. The dynamic of both the 50 and 75 ohm calibration are better than 45 dB for return loss and 1.01 for VSWR over the entire 0 to 6 GHz frequency band. Thus, there is a high-level correlation in terms of reflection due to the connection component between the VNA port and calibration plane.

75 OHM MATCHED LOAD MEASUREMENT

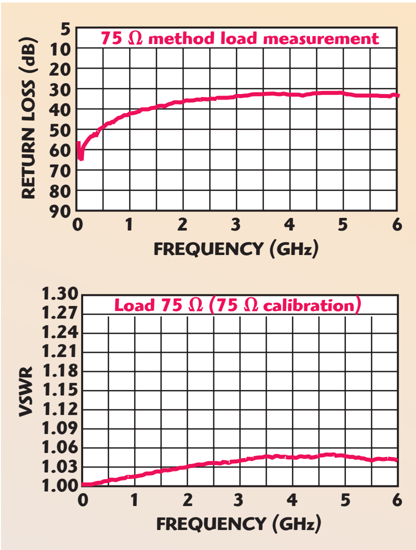
The 75 ohm calibration kit was then used to measure a 75 ohm matched load. The return loss and VSWR results are shown in ***Figure 2***. The VSWR of the matched load is lower than 1.1 in the frequency band from 0 to 6 GHz. These results show that the calibration done with the 75 ohm calibration kit is suitable for measurements up to 6 GHz.

50 AND 75 OHM AIR LINE MEASUREMENTS

Using the two calibrations, Radiall measured 50 and 75 ohm air lines. Since air lines are a perfect coaxial line for measurements by minimizing the effects of a plastic dielectric, it can

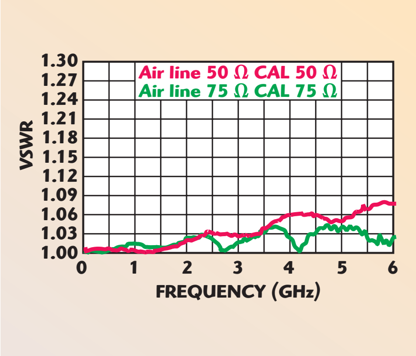


▲ *Fig. 1 Calibration measurement dynamic.*

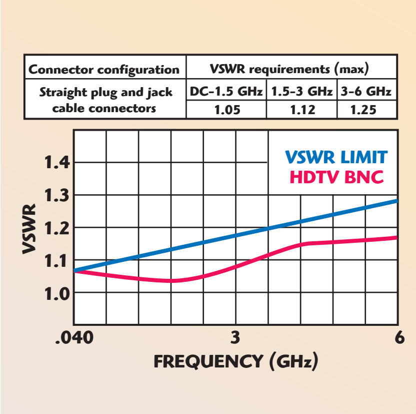


▲ *Fig. 2 Measurement of 75 Ω matched load with 75 Ω calibration kit.*

again correlate the results. Taking into account geometrical dispersion between 50 and 75 ohm air lines and measurement dispersions, the VSWR of the air lines are lower than 1.1, as shown in ***Figure 3***. The conclusion can be made that the measured VSWRs are comparable. The two lines have a VSWR lower than 1.1 up to 6 GHz. Equally important, more than 99.8 percent of the energy passes through the connector, with a low 0.2 percent reflected.



▲ *Fig. 3 50 and 75 Ω air line VSWR measurements by using 50 and 75 Ω calibration kit.*



▲ *Fig. 4 VSWR measurement of the Radiall HDTV to 6 GHz.*

To validate that the 75 ohm calibration presents the same reference (calibration) plane as the 50 ohm calibration, Radiall measured the phase for the two air lines. The maximum phase difference between the two air lines is less than 2 degrees at 6 GHz. Both calibration kits measured with the same phase.

As a result of the process described here, Radiall is able to test and characterize the new BNC connector design through 6 GHz. ***Figure 4*** shows that the new design provides considerable headroom in meeting design goals and ensuring the high-performance needed for evolving broadcast applications.

By testing to see the correlation between an 18 GHz, 50 ohm calibration kit and a 3 GHz, 75 ohm kit, Radiall is able to confidently validate the performance of the 75 ohm kit to 6 GHz. The 75 ohm calibration kit allows the VNA to be calibrated for testing to 6 GHz with accurate results. In fact, the test results will be worst case; products tested will perform better than the measured values.

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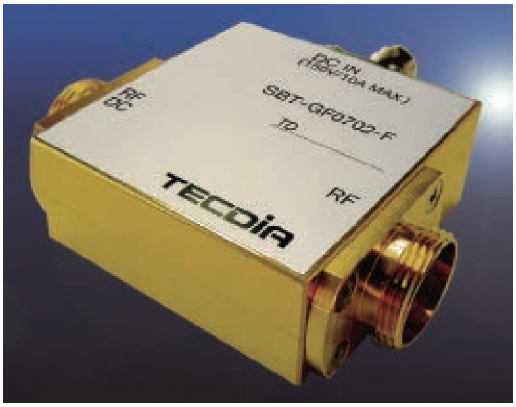
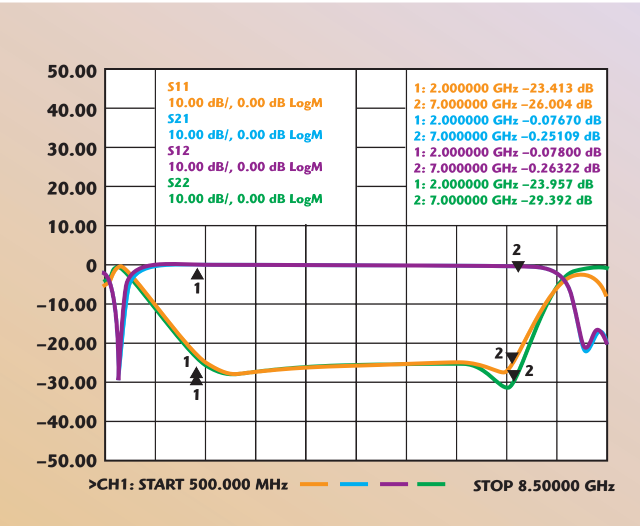


BIAS-T UP TO 100 W AND 7 GHZ FOR GAN

Tecdia specializes in manufacturing high-current bias-Ts with low insertion loss and low return loss. With the increased activity in GaN devices and technologies, Tecdia has developed the SBT-GF0702-F bias-T for high power GaN devices. The SBT-GF0702-F includes coaxial bias-T devices with power handling of 50 W (bias current 20 A, bias voltage 30 V) or 100 W (bias current 10 A, bias voltage 150 V) covering frequencies from 2 to 7 GHz with a return loss of 20 dB minimum and insertion loss of 0.5 maximum. ***Figure 1*** shows the S-parameters versus frequency for the SBT-GF0702-F; ***Table 1*** summarizes its specifications.



▲ *Fig. 1 S-parameters for SBT-GF0702-F.*

The SBT-GF0702-F is housed in a connectorized unit measuring 50 × 52 × 20 mm (excluding connectors) with an APC-7 RF connector and BNC-R (female) DC connector. ***Figure 2*** shows the dimensions of the unit that weighs 200 g.

Tecdia also manufactures bias-T products for GaN applications, including a 10 W (TBT-H06M20-F) and 25 W (TBT-06M20-F) device covering 20 MHz to about 6 GHz in connectorized units along with the SMBT-

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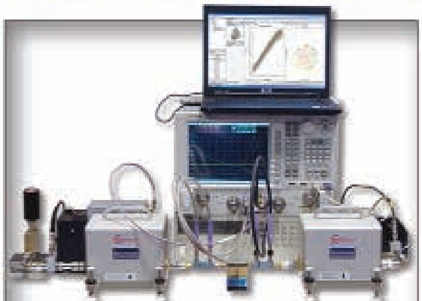

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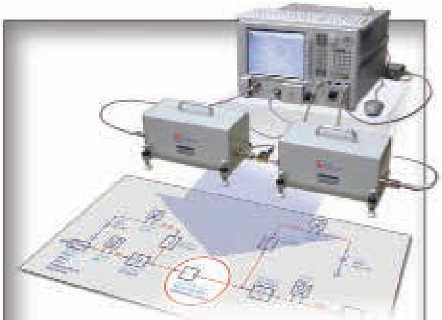

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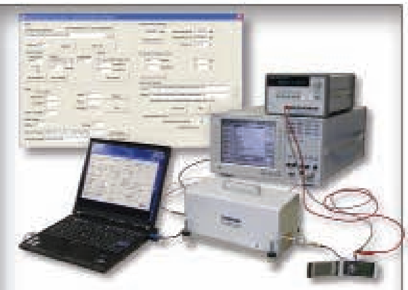

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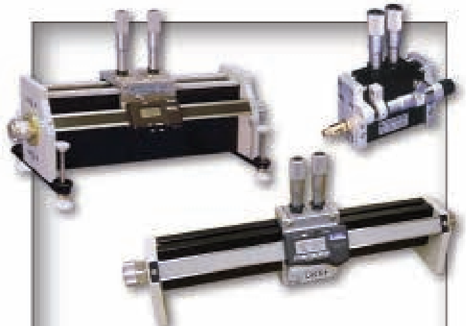

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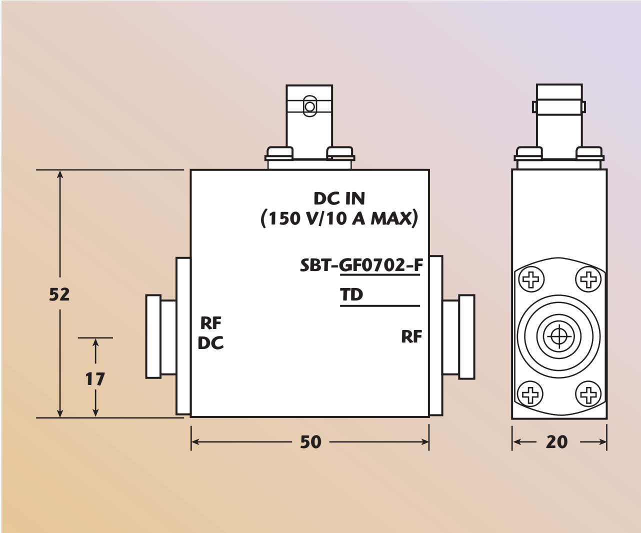
TABLE I

SBT-GF0702-F SPECIFICATIONS

		Bias-T Model	
		SBT-GF0702	
Frequency Range		2~7 GHz	
VSWR (Return Loss)		1.22 max. (20 dB min.)	
Insertion Loss		0.5 dB max.	
Connectors	RF	APC-7	
	DC	BNC-R (Female)	
RF Power		50 W max.	100 W max.
Bias Current		20 A max.	10 A max.
Bias Voltage		30 V max.	150 V max.
Dimensions*		50 × 52 × 20 mm	
Weight		200 g	

*Excluding Connectors



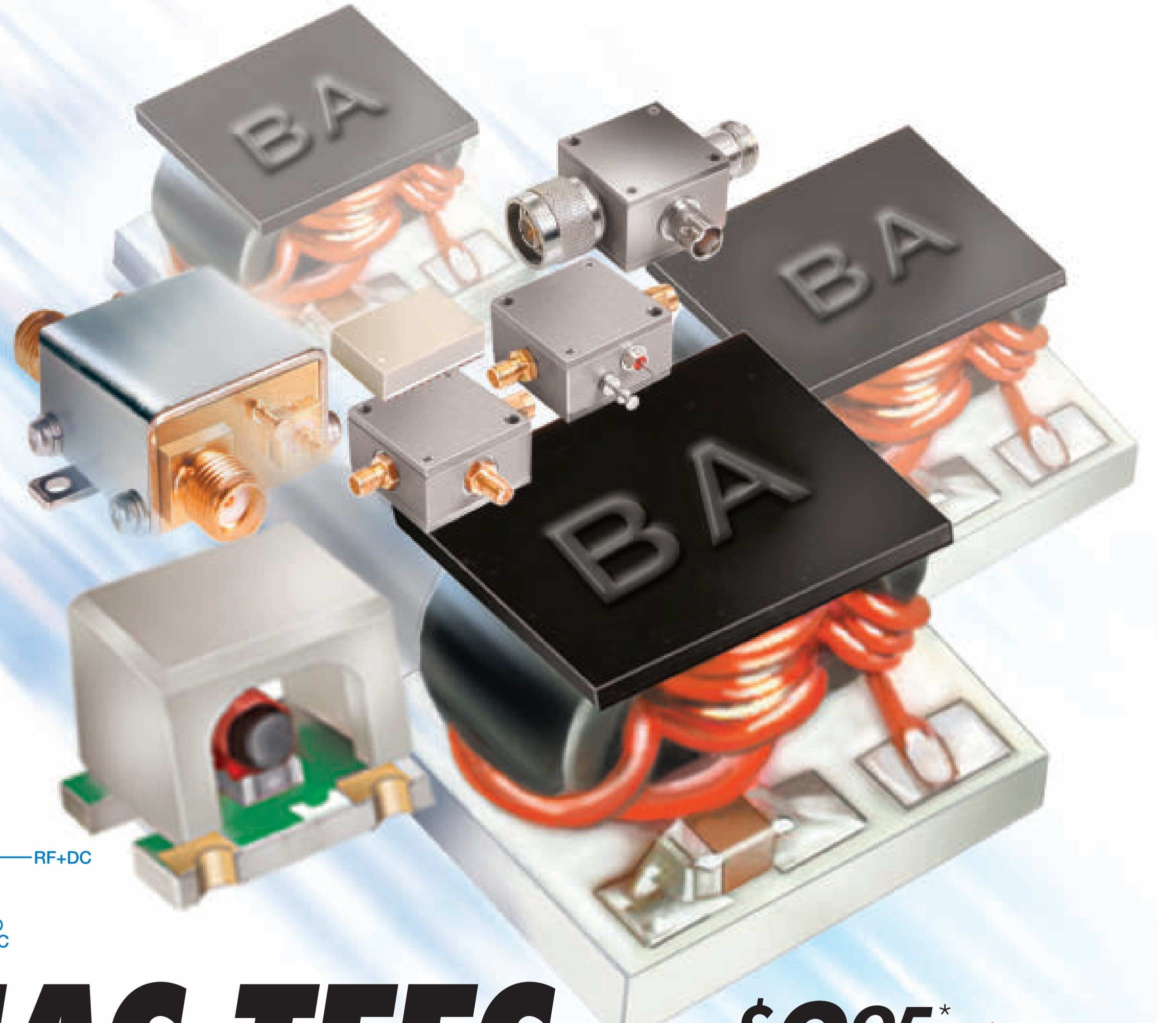
▲ *Fig. 2 Outline drawing for SBT-GF0702-F.*

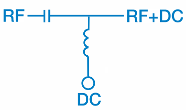
06M20L(or R)-F surface-mount bias-T covering 20 MHz to about 6 GHz and up to 5 W.

Tecdia's model TBT-H06M20 bias-T was selected by Cree for performance data tests of its GaN HEMT MMIC 25 W amplifiers. It was designed to meet the high power and wide bandwidth (20 MHz to 6 GHz) of Cree's model CMPA0060025F.

Tecdia,
Santa Clara, CA
(408) 748-0100,
www.tecdia.com.

RS No. 302





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TYPICAL SPECIFICATIONS

Model	Freq (MHz)	Insertion Loss (dB)	Isolation (dB)	Max Current mA	Price $ea. Qty.10
TCBT-2R5G+	20-2500	0.35	44	200	6.95*
TCBT-6G+	50-6000	0.7	28	200	9.95
TCBT-14+	10-10,000	0.35	33	200	8.45
TCBT: LTCC, Actual Size .15" x .15", U.S. Patent 7,012,486.					**Qty.1-9**
JEBT-4R2G+	10-4200	0.6	40	500	39.95
JEBT-4R2GW+	0.1-4200	0.6	40	500	59.95
PBTC-1G+	10-1000	0.3	33	500	25.95
PBTC-3G+	10-3000	0.3	30	500	35.95
PBTC-1GW+	0.1-1000	0.3	33	500	35.95
PBTC-3GW+	0.1-3000	0.3	30	500	46.95
ZFBT-4R2G+	10-4200	0.6	40	500	59.95
ZFBT-6G+	10-6000	0.6	40	500	79.95
ZFBT-4R2GW+	0.1-4200	0.6	40	500	79.95
ZFBT-6GW+	0.1-6000	0.6	40	500	89.95
ZFBT-4R2G-FT+	10-4200	0.6	N/A	500	59.95
ZFBT-6G-FT+	10-6000	0.6	N/A	500	79.95
ZFBT-4R2GW-FT+	0.1-4200	0.6	N/A	500	79.95
ZFBT-6GW-FT+	0.1-6000	0.6	N/A	500	89.95
ZFBT-282-1.5A+	10-2800	0.6	45	1500	56.95
ZFBT-352-FT+	30-3500	0.4	23	4000	48.95
ZNBT-60-1W+	2.5-6000	0.6	45	500	82.95
ZX85-12G+	0.2-12000	0.6	N/A	400	99.95

ZX85: U.S. Patent 6,790,049.

Note: Isolation dB applies to DC to (RF) and DC to (RF+DC) ports.

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TRIPLE MODE FILTER FOR WIRELESS COMMUNICATION SYSTEMS

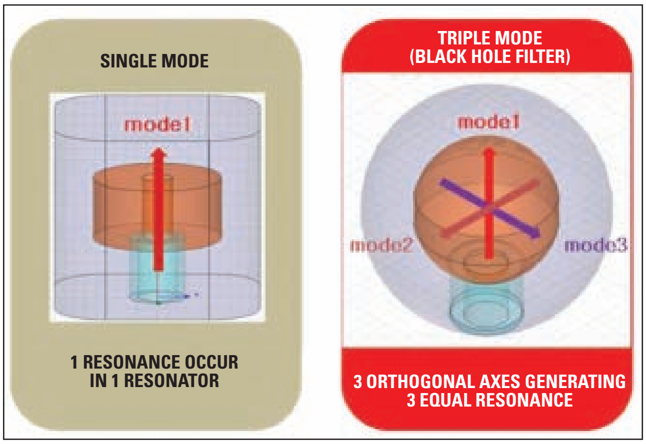
Dale Wildes, Executive Vice President of KMW Communications Inc.

Visit www.mwjournal.com to read this in-depth interview.

Executive Interviews

Cavity filters and dielectric resonator (DR) filters are widely used in communication systems. However, the need for more efficient filters is even greater with the advent of new 4G networks overlaying the existing 3G infrastructure, particularly in areas of limited spectrum and base station resources. Requirements for high attenuation with low insertion loss and small size are key issues in filter technology for communication systems. Many other companies have endeavored to develop dual mode and triple mode filters to fulfill these requirements.

An alternative solution has been proposed by utilizing superconductor technology. But the disadvantages of size, cost, weight and maintenance, not to mention the power required to cool it make superconductor filters impractical. KMW has recently introduced the "Black Hole Filter (BHF)," using the first commercialized "Triple-Mode" technology, which is capable of providing high stop band attenuation with low pass band insertion loss in a relatively small package.



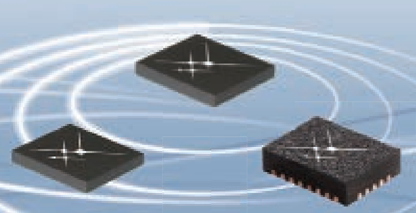
▲ *Fig. 1 Single mode vs. triple mode filter.*

BLACK HOLE FILTER'S TRIPLE MODE RESONANCE

"Triple-Mode" technology enables three distinctive resonances with only one resonant cavity, utilizing a dielectric resonator with high Q value inside the single pocket. Compared to conventional DR filters, the triple mode "Black Hole Filter" offers very low insertion loss, extremely high attenuation, with a significant size reduction. Inside the pocket, waves can travel without bending and distortion, achieving the highest possible Q and steepest band-edge skirts ever in a single pocket filter.

The conventional single mode filter shown in ***Figure 1*** provides only a single resonance mode in the single pocket. The triple mode filter generates three distinctive resonances using the TE01δ mode as the same as the single mode filter. It also enables two additional TE01δ modes allowing the single cavity to function as

KMW USA INC.
Fullerton, CA



Antenna Switches: Discretes and Modules

For 3G/4G Wireless Broadband and Mobile Handset Applications

SP8T–SP10T Antenna Switch Modules—With GSM Harmonic Filter

Part Number / Description	Linear TRx Ports	Rx Ports	Rx SAW	Typ. IL TRx/Tx @ 2.0 GHz (dB)	Typ. Isolation (dB)	IMD (dBm)	H2/H3 Typ. (dBm)	Package (mm)
SKY14152 SP8T, DCS Rx SAW filter	5	1	Yes	-0.6/-1.1	-30	-103	-40/-45	MCM 20L 3.2 x 3.2
SKY18106-455LF SP8T, SPI DigRF interface	6	0	No	-0.6/-0.9	-30	-106	-50/-45	QFN 26L 3.0 x 3.8
SKY18107-455LF SP8T, SPI MIPI interface	6	0	No	-0.6/-0.9	-30	-106	-50/-45	QFN 26L 3.0 x 3.8
SKY18110 SP8T, GPIO logic	6	0	No	-0.8/-0.9	>35	<-102	-45/-42	MCM 20L 3.2 x 3.2
SKY18116 SP8T, small form factor	6	0	No	-0.8/-0.9	>35	<-102	-45/-42	MCM 18L 3.2 x 2.5
SKY18108 SP9T, WCDMA and TD-SCDMA	3	4	No	-0.8/-0.9	>35	<-102	-45/-42	MCM 20L 3.2 x 2.5
SKY13364-389LF SP10T	4	4	No	-0.6/-1.0	>35	<-105	-40/-40	QFN 26L 3.0 x 3.8
SKY13362-389LF SP10T	5	3	No	-0.7/-1.1	>35	<-105	-40/-40	QFN 26L 3.0 x 3.8
SKY13378-389LF SP10T, MIPI interface	5	3	No	-0.7/-1.1	>35	<-105	-40/-40	QFN 26L 3.0 x 3.8
SKY18118 SP10T Rx FEM	5	3	Yes	-0.8/-0.9	>35	<-102	-45/-42	MCM 20L 3.5 x 3.2

SP2T–SP5T Discrete Switches—Suitable for Primary or Diversity Path Applications

Part Number / Description	Linear TRx Ports	Rx Ports	Rx SAW	Typ. IL TRx/Tx @ 2.0 GHz (dB)	Typ. Isolation (dB)	IMD (dBm)	H2/H3 Typ. (dBm)	Package (mm)
SKY13374-397LF SP2T	2	–	No	0.5	>28	–	-50/-50	QFN 16L 2.0 x 2.0
SKY13373-460LF SP3T	3	–	No	0.5	>23	–	-50/-50	QFN 12L 2.0 x 2.0
SKY14155-368LF DP4T, 3G band switch	4	–	No	0.4	>32	–	-50/-50	QFN 12L 2.0 x 2.0
SKY14151-350LF SP4T	4	–	No	0.4	>23	<-115	-45/-45	QFN 16L 3.0 x 3.0
SKY18105 SP4T, Rx diversity, FEM	0	4	Yes	-2.1/-3.3	Filtered	-100	–	MCM 18L 4.5 x 4.5
SKY13358-388LF SP5T	5	–	No	0.7	>27	<-102	-50/-45	QFN 16L 2.3 x 2.3

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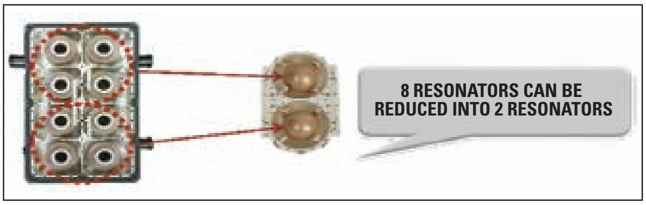
three cavities, which improves the performance and lowers the manufacturing cost. ***Figure 2*** compares the size of the single mode and triple mode filter. The single mode filter with eight resonators can be replaced by the triple mode filter comprised of only two resonators while achieving the same performance, occupying less than a third of the size of the conventional cavity filter.



▲ *Fig. 2 Single mode filter with eight resonators and triple mode filter with two resonators.*

With the deployment of micro base stations including Remote Radio Head (RRH) systems, and the overlay of multiple frequency bands for multiple standards on a single cellular site, the space available for the RF hardware is decreasing. Since RF filters typically occupy a significant fraction of the base station volume, base station manufacturers are looking for filter technologies that offer size and cost reduction, while still meeting the stringent base station RF specifications of insertion loss, rejection and power handling. The multi-mode filter is now an attractive alternative featuring improvements in all these categories.

The differences of the KMW triple mode filter over previous others include:

Previous Triple Mode Filters

- Use DR structure for coupling (limit of structure)
- Difficult to achieve TE01δ mode in applications (limit of function)
- Tuning screw required for tuning (limit of production)
- Difficult to achieve notch filter response (limit of application)

KMW's Triple Mode Filter

- Uses coupling network for coupling (no structural restriction)
- Simple dielectric structure (low cost)
- No tuning screw required for tuning (easy production)
- Notches can be implemented (free application)
- Simple inside structure (simple structure)

Figure 3 shows the single pocket performance achieved by the triple mode BHF in the 800 MHz band for Receive Channel blocking to prevent high power signals from adjacent channel. 20 dB rejection at 1.5 MHz offset from the band edge can be provided by the single pocket BHF. The cascaded BHF pockets can provide more rejection should the application require it.

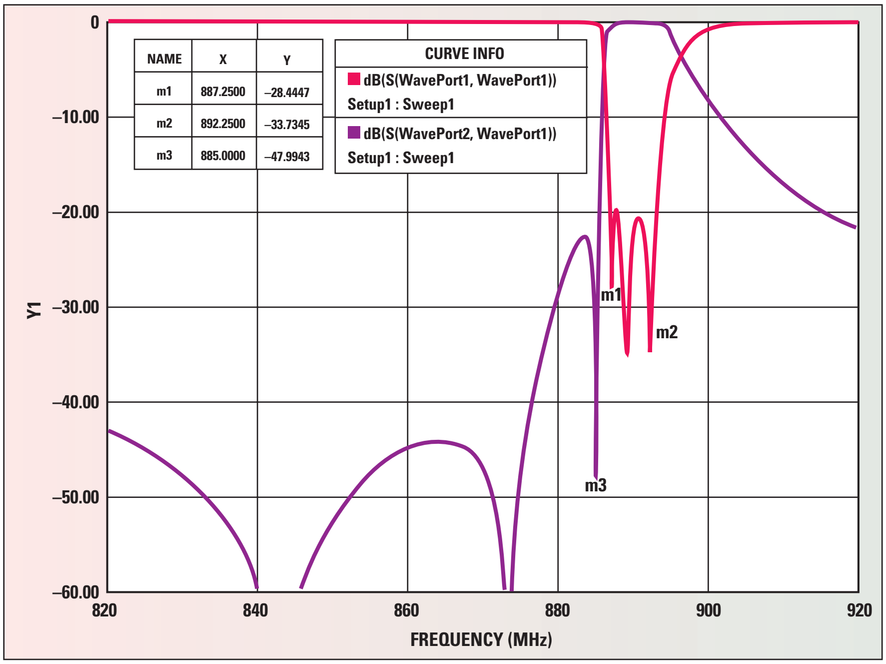
KMW's BHF technology enables a size reduction of about 1/3, and much higher Q value compared to other conventional filters (more than 9 times the Q of cavity filters and 1.6 times that of superconductor filters). Comparative Q values featuring these three filter technologies are shown in ***Table 1***, based on an estimate of a reference 90 × 90 × 90 mm cavity in the 860 MHz band. These advantages are demonstrated by the triple mode resonance and triple notches generated in this single pocket.

KEY APPLICATIONS FOR THE BLACK HOLE FILTER

Some examples of key applications that may be addressed by the BHF are in wireless communication systems where Tx spurious suppression is necessary to meet emission mask requirements, Rx blocking in an interference environment to protect receivers, and co-siting solutions to combine different services in same RF line for reduced CAPEX and OPEX.

700 MHZ/UPPER D BLOCK AND PUBLIC SAFETY BROADBAND

The FCC revised the 700 MHz frequency band plan and service rules to promote the creation of a nationwide interoperable broadband network for public safety and to increase the availability of new and innovative wireless broadband services for consumers. The Commission designated the lower half of the 700 MHz Public Safety Band (763 to 768/793 to 798 MHz) for broadband communications. The FCC also consolidated existing narrowband allocations to the upper half of the 700 MHz Public Safety block (769 to 775/799 to 805 MHz). To minimize interference between broad-



▲ *Fig. 3 Single pocket performance of the Black Hole Filter in the 800 MHz band.*

TABLE I
COMPARISON OF THE BHF AND CONVENTIONAL FILTERS (FREQUENCY=860 MHz, CAVITY SIZE 90×90×90 mm)

	Cavity Filter (TEM mode)	DR Filter (TM mode)	DR Filter (TEσ01 mode)	Super Conductor	Black Hole Filter
Q Factor	7500	11000	18000	45000	24000 × 3 = 72000
Size	▲	▲	▲	▲	❍
Weight	▲	▲	✗	✗	❍
Insertion Loss	✗	▲	●	❍	❍
Attenuation	✗	▲	●	❍	❍
Reliability	●	●	●	✗	●
PIMD	❍	●	●	—	●
Cost	❍	●	▲	✗	●
Handling Power	●	●	●	✗	●

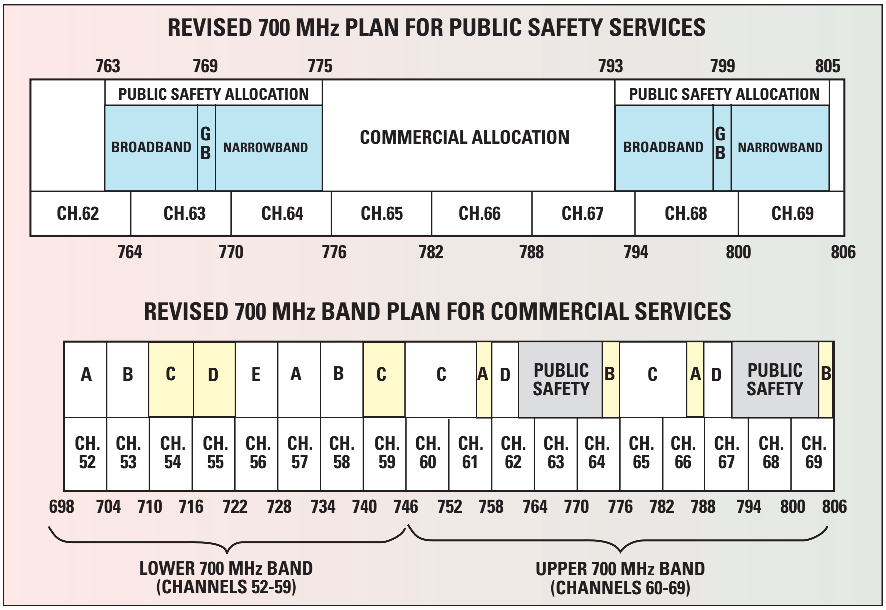
Bad: ✗ Normal: ▲ Good: ● Very Good: ❍

band and narrowband operations, the Commission adopted a one megahertz guard band (768 to 769/798 to 799 MHz) between the public safety broadband and narrowband segments (see ***Figure 4***). The FCC requires a strict emission mask of 76+10log(P) at 1 MHz offset from the Public Safety Broadband channel band edge to protect Public Safety Narrow Band channel integrity.

Table 2 tabulates estimated signal and attenuation levels necessary to meet the 76+10log(P) FCC requirements. According to these estimates, with a single tone at the band edge of 46.5 dBm, the filter should provide more than 40 dB attenuation at 1 MHz offset from that band edge. ***Figure 5***

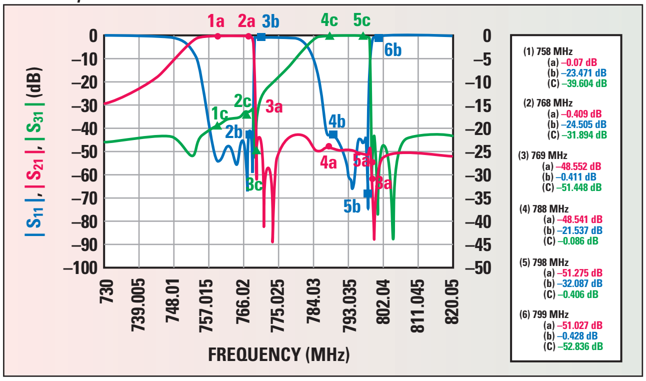
TABLE II
FILTER REQUIREMENTS BY FCC PART 27.53

System Ouput Power	45.0	(W)
	46.5	(dBm)
CH BW (MHz)	5.0	(MHz)
	39.5	(dBm/MHz)
PA ACLR	45.0	(dB)
	-5.5	(dBm/MHz) at PA output's band edge
	40.5	(dB) rejection by filter @ 1 MHz offset to meet -46 dBm/MHz req. (76+10logP)



▲ *Fig. 4 Revised 700 MHz band plan.*

presents the simulation results of a duplexer of the 700 MHz upper D and Public Safety Broadband satisfying the FCC regulation by applying a BHF with 0.6 dB max insertion loss. It shows -48 dB at the Public Safety Narrow Band down link starting point of 769 MHz and -51 dB at 799 MHz at same point on the up link side.



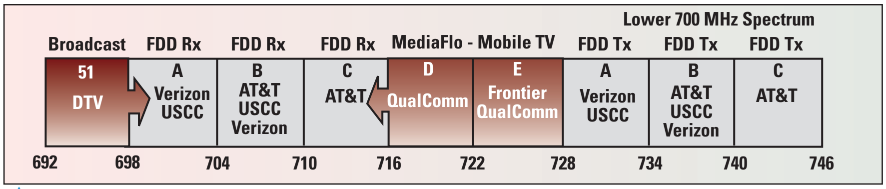
▲ *Fig. 5 700 MHz duplexer for upper D and Public Safety Broadband.*

LOWER 700 MHZ RX BLOCKING FILTER TO PROTECT FROM THE DTV AND MEDIAFLO

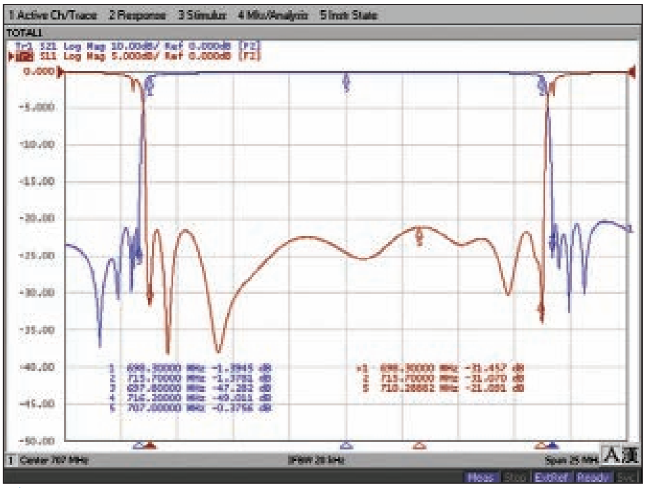
Another useful application is Rx blocking to prevent high power signal leakage from DTV channel 51 and MediaFlo on the lower 700 MHz band, as shown in ***Figure 6***. New LTE systems will be deployed in lower 700 MHz band A, B and C Block. The high power signal from adjacent broadcasting systems will impact the LTE receiver and significantly degrade system performance. To avoid receiver performance degradation, the system needs to employ a filter with improved out of band rejection. The wall type filter shown in ***Figure 7*** produces more than 40 dB attenuation at 500 kHz offset. The BHF filter appears to be the best solution for Rx blocking in this high-interference environment.

CO-SITING SOLUTIONS

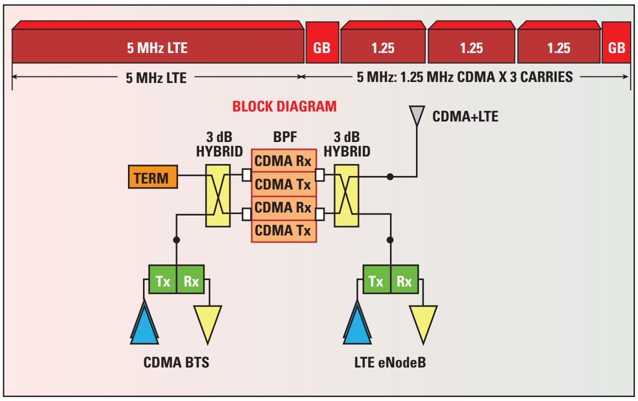
2G, 3G and 4G LTE/M-WiMAX technologies are expected to co-exist, installing in same sites and sharing antennas. This implies that co-siting solutions will be required for CAPEX and OPEX. ***Figure 8*** shows a typical example of an in-band combiner to support that of three carriers, CDMA and the new 5 MHz LTE that must be



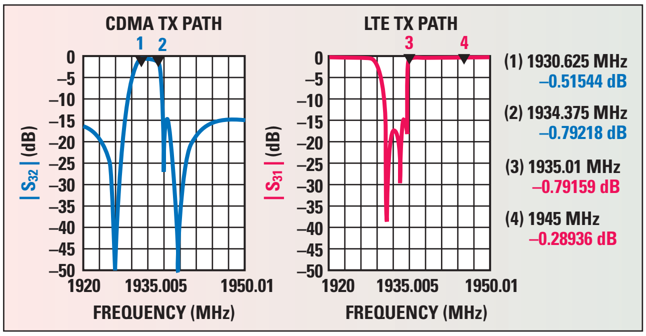
▲ *Fig. 6 Lower 700 MHz spectrum.*



▲ *Fig. 7 700 MHz Wall Type filter to protect from DTV and MediaFlo.*



▲ *Fig. 8 Co-siting example of CDMA plus LTE.*



▲ *Fig. 9 Filter response of co-siting combiner (CIB).*

combined over 10 MHz of bandwidth with a narrow guard band. Conventional combiners require more than 5 MHz for guard band, but the Constant Impedance BPF (CIB) implemented by the BHF only needs 625 kHz guard band to combine two different services. The simulation results are presented in ***Figure 9***.

KMW has been bringing industry leading filter technology and RF products to market since the early days of CDMA in 1995. KMW enables a new world of filter technology "Black Hole Filter" by the "Triple Mode" resonance and offers solutions that are aimed at maximizing spectrum utilization and reducing size by market demands.

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RS No. 301

COMPACT, LOW SSB PHASE NOISE WIDEBAND VCOs COVER 6 TO 20 GHz

Hittite has introduced two wideband voltage-controlled oscillators (VCO): The HMC732LC4B and HMC733LC4B, which operate between 6 to 12 and 10 to 20 GHz, respectively. While complementing and extending the frequency range of Hittite's GaAs HBT wideband VCO offerings, they are ideal for military radar and EW & ECM, as well as test and measurement.

The HMC732LC4B and HMC-733LC4B integrate a resonator, negative resistance device and varactor diode in a single 4 × 4 mm RoHS compliant SMT package. The Vtune port of the VCOs accepts an analog tuning voltage from 0 to +23 V and accommodates a very fast tuning bandwidth. The HMC732LC4B has SSB phase noise of -65 dBc/Hz at 10 kHz and -95 dBc/Hz at 100 kHz, whereas the HMC733LC4B has SSB phase noise of -60 dBc/Hz and -90 dBc/Hz at 10 and 100 kHz, respectively. Both provide a single ended RF output matched to 50 ohms.

The HMC732LC4B and HMC-733LC4B combine the attributes of low phase noise, compact size and wide tuning range, making them ideal for applications where tuning range and high performance are of primary importance. The HMC-732LC4B and HMC733LC4B operate from a single +5 V power supply, while drawing 54 and 70 mA of current, respectively. Their SMT footprint requires up to 90 percent less PC board area when compared to traditional MCM and discrete hybrid VCO approaches.

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

RS No. 303

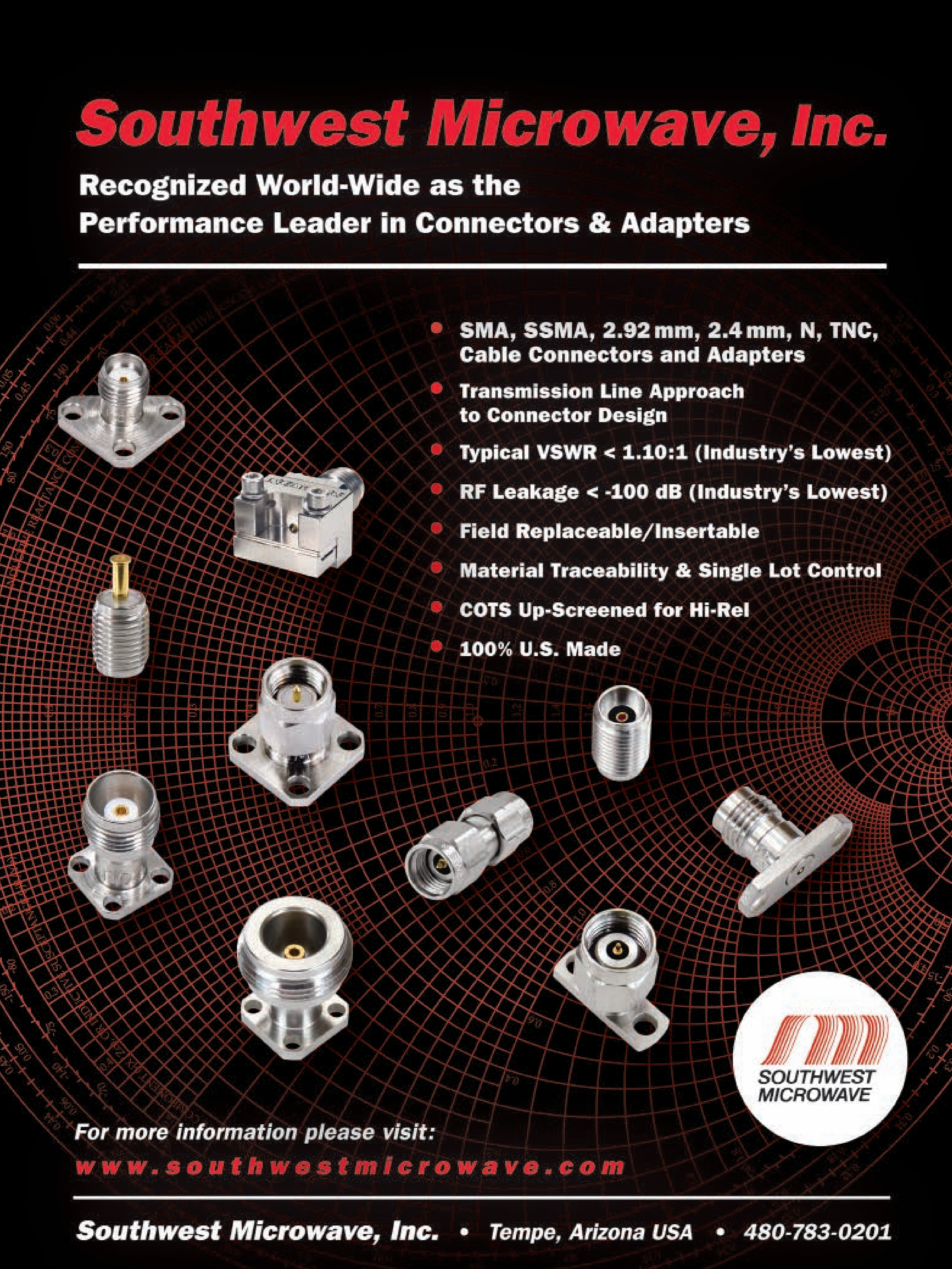
Power Divider/Combiners Offer Low-loss and High Isolation

A microwave divider/combiner is a passive reciprocal RF/microwave component responsible for the vector summation of two or more signals. As such, they play an important role in the distribution of signals through higher-level systems such as mobile base stations, distributed antenna systems, in-building wireless systems and cable distribution networks. Increasingly, these systems must support more signals and greater bandwidths, requiring new levels of performance. Additionally, the move toward remote radio heads is also placing new constraints on cost, size, weight and specifications for stringent outdoor use.

To address these needs, MECA Electronics has implemented its V-Line power divider/combiner product family based on microstrip techniques using high quality materials to lower insertion loss. The V-line compact power divider/combiner offers increased power rating and extended frequency range. These compact, microstrip divider/combiners provide minimal insertion loss while delivering high isolation between output ports with exceptional phase and amplitude balance. These two- through 16-way, 40 W power divider/combiners are optimized for excellent performance across all wireless bands from 0.7 to 2.7 GHz. The components, which are available from stock within two weeks ARO, are available in N, SMA, BNC or TNC connector configurations. The two-way 800 series is a divider/combiner that exhibits insertion loss below 0.5 dB with 27 dB typical isolation. All units are rated for maximum input power as a divider or balanced combiner with load VSWR of 1.20:1 or better. All output/power combiner ports are in-phase (0° difference) and are rated for an operating temperature from -55° to +85°C. They can be manufactured to IP65 weather proof specifications for outdoor (tower mounted) applications.

MECA Electronics, Denville, NJ (866) 444-6322, www.e-meca.com.

RS No. 306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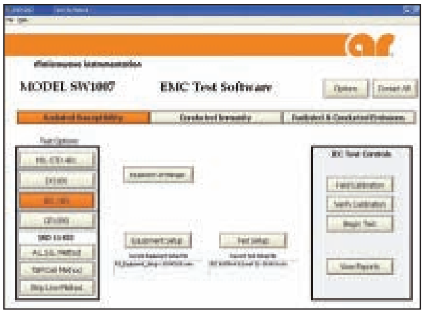
SYSTEM SIMULATION SOFTWARE

ACS has recently released a new version of the LINC2 VSA (Visual System Architect) system simulation software. Version 1.09 adds new system component models to the VSA's components menu. This new version of the VSA offers a new file-based multi-tone signal source for stimulation of system schematics with multiple simultaneous input tones. The associated signal file can arbitrarily specify the signal amplitudes, frequencies and signal phases for each signal in the input signal set. The Visual System Architect offers the flexibility and ease of use of schematic-based RF system simulation combined with a comprehensive array of analysis methods and graphic displays for designing at the system level.

Applied Computational Sciences (ACS) LLC,
Escondido, CA (760) 612-6988,
www.appliedmicrowave.com.
RS No. 310



SOFTWARE FOR COMPLIANT TESTING

AR's SW1007 software is a standalone program that combines conducted immunity test software and radiated susceptibility test software into one user-friendly package suitable for corporate to professional test lab users. The software automatically performs both calibration and immunity testing in full compliance with IEC 61000-4-3, 4-6; MIL STD 461/462 RS103, CS114 and RTCA/DO160 Section 20 specifications. The new version has an updated user interface including a tab system and organizes all the features for quick, easy access and makes selecting test standards much easier. The SW1007 also has the ability to control more equipment and the report generating feature has been enhanced to offer more control and customization.

AR RF/Microwave Instrumentation,
Souderton, PA (215) 723-8181, www.ar-worldwide.com.
RS No. 311

USB DESIGNER'S KIT

The HMC-DK008 Serial/Parallel USB Interface Designer's kit has been expanded to provide a user friendly interface for programming Hittite's family of interface driver/controllers, digital attenuators and variable gain amplifiers. This kit allows the designer to set desired attenuation and gain states, toggle between serial and parallel control modes, and construct custom serially clocked input signals. The HMC-DK008 Designer's Kit includes a Serial/Parallel USB Interface Board, custom USB and ribbon cable assemblies, and software CD-ROM.

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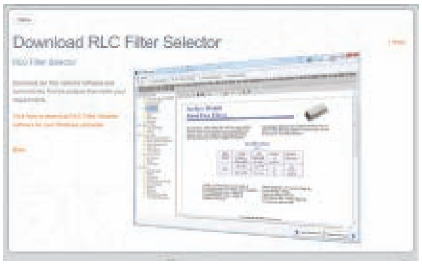
www.hittite.com.
RS No. 312



FILTER SELECTOR

RLC Electronics, a supplier of RF and microwave components, is pleased to announce and introduce the company's new website, www.rlcelectronics.com. The site now includes a new search capability and a downloadable filter selector software and performance program. Download the company's filter selector software and automatically find the product that meets your requirements.

RLC Electronics Inc.,
Mount Kisco, NY (914) 241-1334,
www.rlcelectronics.com.
RS No. 313

Three words you've never heard from us: "Sorry, that's obsolete."

At Aeroflex / Metelics our commitment is to support your program needs for as long as you need your devices. Today, we call it Product Lifeline Assurance. But for over 30 years, it's just how we've been doing business. While our competitors regularly change manufacturing strategies and discontinue products—forcing you into unscheduled and unfunded redesigns—we remain committed.

In short, we promise:

- **To stand behind our policy not to obsolete products.**
- **To be the benchmark for consistency in your supply chain strategy.**
- **Wherever possible, to replace competitors' obsolete devices—offering assurance on these devices, too.**

Product Lifeline Assurance. Support you can rely on for your high performance diodes, passive and integrated devices, and custom semiconductors.

www.aeroflex.com/metelicsPLA

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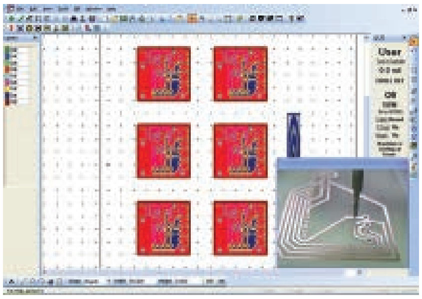


ISOPRO 3.0® SOFTWARE

T-Tech Inc. announces the release of IsoPro® 3.0, Mill Path Generation and CAD editing software for Quick Circuit users. The newly added features are "CheckByTouch®" broken bit detection, "ContactByTouch®" automatic depth sensing, auto tool change, tool management including Linking, improved mill/drill path optimization, and new multilingual help files with embedded graphics. T-Tech offers a new Live Machine Status Bar, with added mouse functions and much more. IsoPro 3.0 exceeds anything on the market today, offering upgrades to existing customers. IsoPro 3.0 is Windows 2000 through Win 7 and Mac 10.X compatible. The software was recently released at MilCom 2010.

T-TECH Inc.,
Norcross, GA (770) 455-0676,
www.t-tech.com.
RS No. 314

APPLICATION PROGRAMMING INTERFACE

Wireless Applications Group announced it has beta-released a suite of web service Application Programming Interface (API) for performing telecom infrastructure analysis called the Telecom API Suite. Carriers, tower companies, web app programmers and wireless service providers alike can immediately begin amping up their software with power from the Wireless Applications' cloud service. Anyone can sign up for a free dev key and begin immediately seeing how easy it is to integrate these powerful functions into their site. Included in the site are samples of PHP, C#, Javascript and even Objective C for those popular phone apps. There is even an assortment of simple examples of map integration to get started.

Wireless Applications Corp.,
Bellevue, WA (425) 643-5000,
www.wacorp.net.
RS No. 315

IEEE Radio and Wireless Week

16 - 19 January 2011, Phoenix, AZ, USA

http://www.radiowirelessweek.org/

Register Now! http://www.radiowirelessweek.org/

Start The New Year With A Week of Wireless 16–19 January 2011 in Phoenix, AZ

Join us in Phoenix, Arizona for the 2011 IEEE Radio and Wireless Week (RWW) which now includes three all new conferences. In addition to the Radio and Wireless Symposium (RWS) and IEEE Topical Meeting on Silicon Monolithic Integrated Circuits in RF Systems (SiRF), the IEEE MTTS and RWW have created three new conferences to fill a void in the society's conference offerings. This first being the IEEE Topical Conference on RF/Microwave Power Amplifiers (PAWR), the society's first conference publishing an archival digest dedicated to the topic of power amplifiers. The second new conference is the IEEE Topical Conference on Biomedical Wireless Technologies, Networks & Sensing Systems (BioWireleSS) targeting one of the most exciting and rapidly growing areas due to the potential of wireless medical devices. The third new conference is the IEEE Topical Conference on Wireless Sensors and Sensor Networks (WiSNet), which utilizes technologies from the RWS, PAWR, and SiRF to develop new applications related to BioWireleSS, consumer, commercial, and military sensor markets.

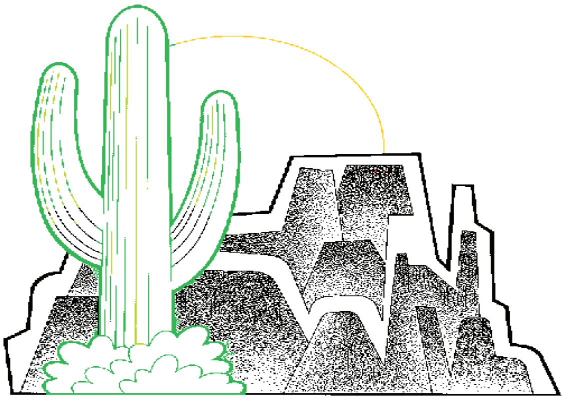
RWW 2011 Highlights

IEEE Radio and Wireless Symposium
IEEE Topical Conference on Wireless Sensors and Sensor Networks **NEW!**
IEEE Topical Conference on Biomedical Wireless Technologies, Networks & Sensing Systems **NEW!**
IEEE Topical Conference on RF/microwave Power Amplifiers **NEW!**
11th Topical Meeting on Silicon Monolithic Integrated Circuits in RF Systems
35 Technical Oral Sessions - Mon - Wed, 17-19, Jan., 2011
Interactive Poster Sessions - Mon/Wed, 17,19, Jan., 2011
Student Paper Competition Finals - Mon, 17, Jan., 2011
Workshops - Sunday afternoon, 16, Jan., 2011
"Inter- and Intra-vehicle Wireless Communications & Networking"
"Wireless Communications for Smart Grid"
"Wireless Biomedical Applications"
"ZigBee Applications"
Panel Sessions
"Ultra-wideband (UWB) Technology: Past, Present, and Future"
"Cell Phone Tower Myths and Misconceptions"
Joint RWW/SiRF Plenary - Tue, 18, Jan., 2011
Joint RWW/SiRF Banquet - Tue, 18, Jan., 2011
Exhibits – Mon-Wed, 17-19, Jan., 2011

Distinguished Lecturer Talks

Inkjet-Printed Paper/Polymer-Based "Green" RFID and Wireless Sensor Nodes: The Final Step to Bridge Cognitive Intelligence, Nanotechnology and RF?
Dr. Manos Tentzeris, GEDC Associate Director for RFID/Sensors Research, *Georgia Institute of Technology*
Relay Node Placement in Wireless Sensor Networks
Prof. Guoliang (Larry) Xue, *Arizona State University*

More to be announced soon!!!



ORAL PRESENTATIONS - POSTERS - WORKSHOPS - PANELS - EXHIBITIONS

HTTP://WWW.RADIOWIRELESSWEEK.ORG

Components

Crystal Notch Filter

VENDORVIEW

Anatech Electronics introduced the AS10CN801 crystal notch filter, which offers a notch center frequency of 10 MHz and notch bandwidth of 1 kHz. The filter's rejection in the notch bandwidth is 80 dB, low-side passband is 9.8999 to 9.999 MHz, high-side passband is 10.001 to 10.201 MHz, and insertion loss is less than 1.3 dB. The filter's power handling ability is 10 dBm, and it employs SMA female connectors. The AS10CN801 measures 6"× 1" × 1".

Anatech Electronics,
Garfield, NJ
(973) 772-4242,
www.amcrf.com.

RS No. 216

RF Coaxial Cable Assemblies

Crystek has ruggedized its LL142 low-loss RF cable assemblies by incorporating a spiraled stainless steel casing, along with extra fortification provided by heavy-duty adhesive strain relief with a Neoprene jacket. This added measure of protection eliminates the failures commonly caused by cable flexion and compression. At 18 GHz, the new Armored LL142 assemblies feature attenuation of 0.36 dB/ft. and VSWR characteristics of < 1.3. Crystek's Armored LL142 Series cables offer shielding effectiveness of greater than -110 dB with an operating temperature range of -55° to +85°C (extended range of -55° to +125°C available through special order).

Crystek Corp.,
Fort Myers, FL (239) 561-3311,
www.crystek.com.

RS No. 223

Ka-band Up/Down Converter

The Ka-band up/down converter assembly is a custom designed system. It operates in an anechoic chamber in order to simulate a radar receiver with realistic target echoes. The system includes four transmitter channels (up converter), one receiver channel (down converter), detectors, control circuitry and DC power suppliers. The system operates at 33.3 to 36.3 GHz RF frequency range and IF of 4.3 to 5.3 GHz. The receiver channel has the gain of greater than 40 dB and noise figure of 5 dB, while the transmitter P1dB varies from +20 to +32 dBm with different channels according to requirements. The LO input is about 13 dBm from 29 to 31 GHz. The LO is multiplied by 2 and then shared by all the channels. The assembly is in the standard rack mountable box with the dimension of 19" × 7" × 19".

Ducommun Technologies Inc.,
Carson, CA (310) 847-2859,
www.ducommun.com.

RS No. 224

Temperature Variable Attenuator

VENDORVIEW

EMC Technology introduces a new wideband surface-mount temperature variable attenuator optimized for performance from DC to 20 GHz. Using EMC's patented Thermopad® technology, the WTVA offers the best performance to date for high frequency applications including optimal temperature coefficients of attenuation (TCA) at frequencies from 12.4 up to 20 GHz. These devices have been noted for their excellent performance and small size, measuring only 0.070" × 0.060". The WTVA wideband temperature variable attenuator is available in a RoHS compliant solder finish with dB values from 2 to 6 dB and negative coefficient slopes from 0.003 to 0.006.

EMC Technology,
Stuart, FL
(772) 286-9300,
www.emc-rflabs.com.

RS No. 217

RMS Power Detector

The LTC5583 is a 40 MHz to 6 GHz dual channel, matched RMS power detector, offering over 55 dB isolation at 2.14 GHz. In RF

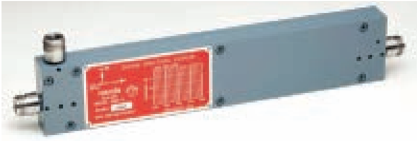
power amplifier (PA) applications, the LTC5583 provides a simple solution for accurately measuring forward power, reverse power and voltage standing wave ratio (VSWR). The device comprises a pair of 60 dB dynamic range RMS detectors that are matched to 1.25 dB. This provides accurate RF power measurement of high crest-factor signals such as those used in LTE, WiMAX, W-CDMA, TD-SCDMA and cdma2000 3G or 4G base stations and other high performance radios employing complex modulation waveforms.

Linear Technology,
Milpitas, CA
(408) 432-1900,
www.linear.com.

RS No. 218

Coaxial Directional Coupler

VENDORVIEW



The model 3000-10 is a coaxial directional coupler that delivers 10 dB of coupling from 225 MHz to 10 GHz, has flat frequency response characteristics, high directivity, and is ruggedized for both commercial and military applications. The model 3000-10 handles 200 W average and 10 kW peak RF power, has directivity of at least 30 dB, insertion loss of 0.7 dB or less, maximum VSWR of 1.15:1, maximum deviation from nominal of ±1.2 dB, absolute calibration accuracy of ±0.1 dB per 10 dB step, weighs 1.8 oz., measures 10.68" × 2.13" × 0.88", and uses Type-N female connectors. The model 3000-10 is available from Narda for immediate delivery.

Narda Microwave-East,
Hauppauge, NY (631) 231-1700,
www.nardamicrowave.com/east.

RS No. 220

Calibration Module

The Calibration Module is a two-port microwave assembly providing TTL selectable thru, open, short or 50 Ω termination load on the input port, and thru or 50 Ω termination on the output port. The four available Input/Output connections are: Short/Term, Open/Term, Term/Term or Thru/Thru. This unit permits calibration of coaxial test cables with far end termination type to be TTL controlled, where those far ends are in physically inconvenient locations such as in vacuum or thermal chambers. This permits cable recalibration without personnel opening or entering the chamber. This can save considerable time in production test environments.

Paciwave Inc.,
Sunnyvale, CA (408) 745-0385,
www.paciwave.com.

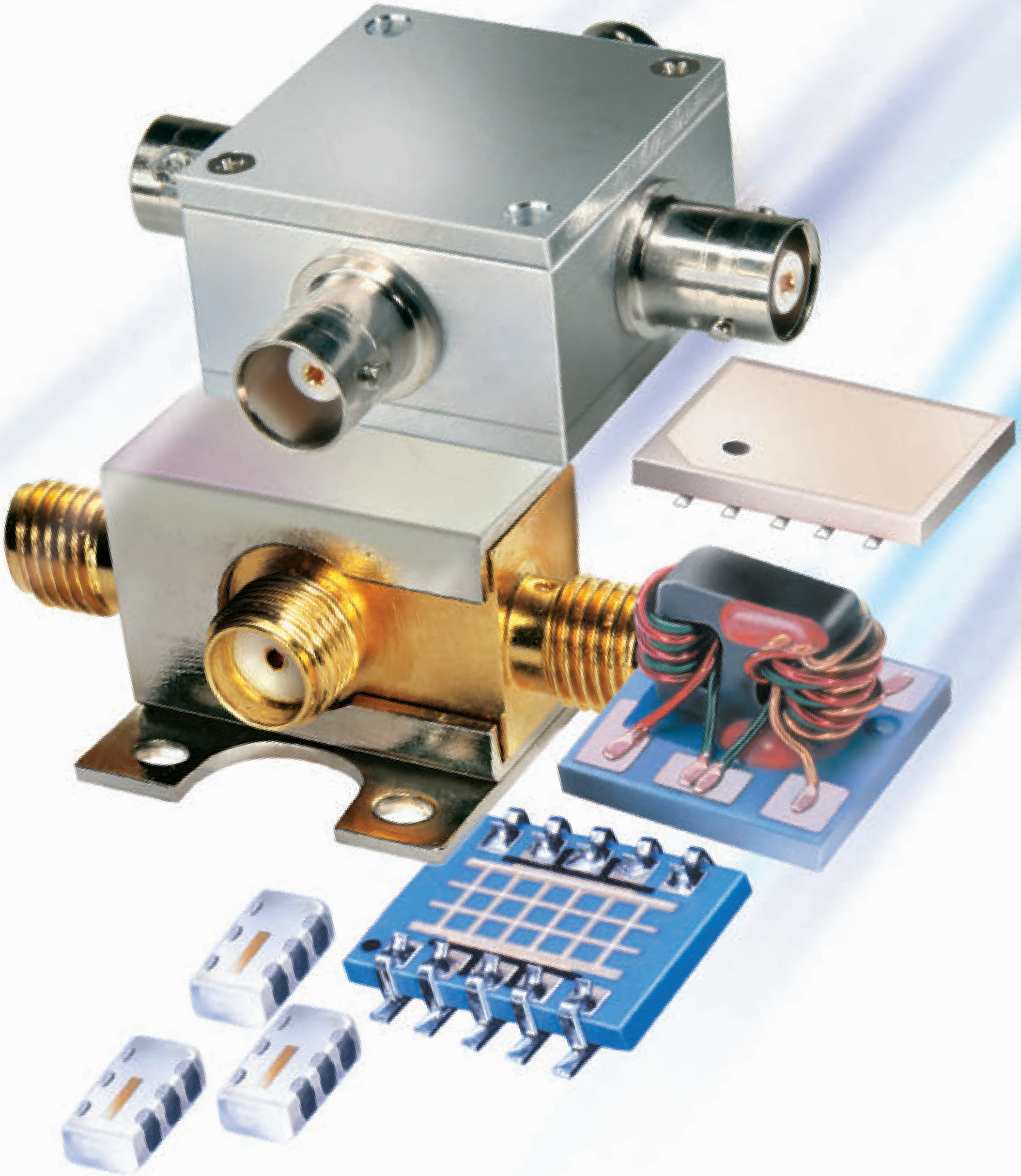
RS No. 227

Two-way Power Divider

PS2-52-450/8S is a two-way power divider that covers the frequency range of 5 to 40 GHz with 2.2 dB insertion loss, 13 dB isolation and 1.90:1 maximum VSWR. Amplitude and phase balance are 0.8 dB and ±10 degrees, respectively. Power rating is 1 W and 2.92 female connectors are utilized.

Pulsar Microwave Corp.,
Clifton, NJ (973) 779-6262,
www.pulsarmicrowave.com.

RS No. 221



Directional/Bi-Directional

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RoHS COMPLIANT

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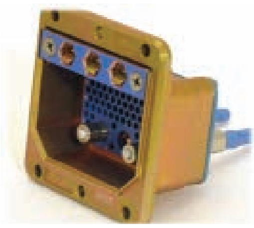
U.S. patent 7739260

IF/RF MICROWAVE COMPONENTS

396 rev M

Optical Adapter



Radiall USA Inc. introduces its new size 5 EN4531 Optical Adapter for existing MIL-C-81659B/DSX series (ARINC 404 standard) connectors. It is available in both male and female versions for either plug or receptacle DSX connectors. It is suitable for standard EN4531 multimode optical contacts (termini). With this new adapter, the optical contact performance meets and exceeds the requirements of the EN4531 standard. The two piece adapter can easily be removed using a standard extraction tool for size 5 coaxial contacts. This new size 5 EN4531 Optical Adapter is a key accessory to allow for fiber optic technology upgrade in existing rack and panel interconnect applications.

Radiall USA Inc.,
Chandler, AZ
(480) 682-9400,
www.radiall.com.

RS No. 228

Air Dielectric Directional Couplers

RLC Electronics' miniature air dielectric directional couplers are rugged lightweight devices that offer lower insertion loss than comparable stripline units. The simplified construction allows for greater flexibility in creating customized configurations. Any port can be used as the input with these symmetrical devices. The standard units are available with a choice of coupling values and frequency ranges and an optional termination.

RLC Electronics Inc.,
Mount Kisco, NY
(914) 241-1334,
www.rlcelectronics.com.

RS No. 222

Amplifiers

GaN Amplifier

The model SSPA 2.8-3.8-200-RM is a high power, CW RF amplifier that operates from 2.5 to 4 GHz in a rack-mounted configuration. This GaN amplifier is easily extendable for operation from 2.5 to 6 GHz with 200 W of RF power across the full band. Also, higher power levels are easily achievable. Please contact the factory. It is packaged in a 3U high, 19-inch rack-mounted enclosure. This amplifier has a minimum saturated output power of 200 W. This amplifier offers a typical saturated gain of 40 dB with a typical power flatness of ±1 dB. Input and output VSWR is 1.5:1 maximum. This RF rack-mounted amplifier operates from 208 to 220 Vac.

Aethercomm,
Carlsbad, CA (760) 208-6002,
www.aethercomm.com.

RS No. 229

X-band Amplifier

AML announces the availability of a low noise X-band amplifier, model number AML812L3003. This LNA operates in the frequency range of 8 to 12 GHz with small-signal gain over 30 dB and a noise figure of 1.1 dB typical. Output P1dB is +10 dBm minimum. This amplifier is available in a SMA connectorized housing with internal voltage regulation and reverse voltage protection.

AML Communications Inc.,
Camarillo, CA (805) 388-1345,
www.amlj.com.

RS No. 230

Solid-state Amplifiers

VENDORVIEW

AR RF/Microwave Instrumentation has introduced a family of new solid-state amplifiers that are more compact, more efficient and more powerful than previous models. The new "S" Series covers 0.8 to 4.2 GHz and powers up to 800 W. These models employ a new design that delivers more than twice the power of older models. With these improvements, AR has maintained the superior rugged design for mismatch tolerance and excellent linearity.

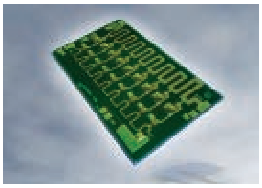
AR RF/Microwave Instrumentation,
Souderton, PA (215) 723-8181,
www.ar-worldwide.com.

RS No. 231

MMIC Distributed PA

VENDORVIEW

The model EWH2001ZZ is a distributed power amplifier (PA) for commercial, industrial and military applications from DC to 20 GHz. Based on GaAs pseudomorphic high-electron-mobility-transistor (PHEMT) technology, the monolithic-microwave-integrated-circuit (MMIC) distributed PA delivers +26 dBm typical output power at 1 dB compression (P1dB) through 12 GHz and typically +23 dBm through 20 GHz. The amplifier features impressive gain and gain flatness over its broad bandwidth, with small-signal gain of typically 19 dB at 2 GHz and typically 18 dB at 20 GHz, with gain flatness of ±2 dB from DC to 20 GHz. The broadband amplifier provides usable gain of 10 dB through 30 GHz.

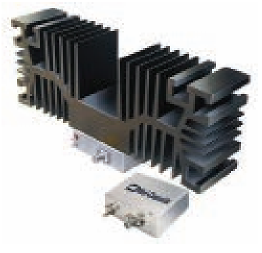


Endwave Corp.,
San Jose, CA
(408) 522-3100,
www.endwave.com.

RS No. 232

Coaxial High Power Amplifier

VENDORVIEW



The ZVE-3W-183+ is a coaxial high power amplifier that operates in a frequency range from 6 to 18 GHz. Features include: high power, 3 W; wideband, 6 to 18 GHz; high IP3, +44 dBm typical; high dynamic range; high gain, 35 dB typical and good directivity, 35 dB typical; and internal voltage regulated for 13 to 18 VDC. Applications include: radar, video and instrumentation, and lab use.

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

RS No. 233

X-band Power Amplifier

VENDORVIEW

A series of compact X-band amplifiers are introduced by MITEQ Inc., covering 30 to 34 dBm. The model AMF-6B-08501070-80-33P-ISO delivers over 33 dBm of power over the band 8.5 to 10.7 GHz, with over 30 dB gain and ±0.75 dB flatness. P1dB is over 34 dBm above 10 GHz. Noise figure is less than 8 dB, port VSWR is less than 1.5:1, and it draws about 1.4 A from a single +12 to +15 V DC supply. Output isolator is optional. Typical output IP3 is over 41 dBm. The housing has a footprint of only 3" by 1.9" and 0.9" high with SMA connectors. Heatsink and other control features are also available as options.

MITEQ Inc.,
Hauppauge, NY (631) 436-7400,
www.miteq.com.

RS No. 234

Successive Detection Log Video Amplifiers

PMI model SDLVA-6G18G-CD-1 Successive Detection Log Video Amplifiers (SDLVA) offers 75 dB dynamic range over the frequency range of 6 to 18 GHz. This model offers an ultra-fast rise time of 10 nsec maximum and a recovery time of less than 60 nsec. The unit is temperature compensated such that log linearity over temperature remains less than ±2.5 dB over the full operating temperature range of -40° to +85°C. This model is supplied in a compact housing measuring only 3.2" × 1.8" × 0.4". Optional frequency ranges covering 100 MHz to 26.5 GHz are available.

Planar Monolithics Industries Inc.,
Frederick, MD
(301) 662-5019,
www.pmi-rf.com.

RS No. 235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www.Spectrum-et.com

Integrated Circuit

I/Q Upconverter/Transmitters

VENDORVIEW

The HMC924LC5 is a GaAs MMIC I/Q upconverter with variable gain that operates from 10 to 16 GHz, delivers a small-signal conversion gain of 15 dB and has -30 dBc of sideband rejection. The HMC924LC5 provides a high output IP3 of +14 dBm while maintaining LO/RF rejection of 15 dBc or better across the operating frequency band. The HMC-925LC5 is a GaAs MMIC I/Q upconverter with variable gain that operates from 5.5 to 8.6 GHz, delivers a small-signal conversion gain of 16.5 dB and has -30 dBc of sideband rejection. The HMC925LC5 provides a high output IP3 of +29 dBm while maintaining LO/RF rejection of 22 dBc or better across the operating frequency band.

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

RS No. 237

Sources

Ku-band Synthesizer

The THOR-16000 Ku-band frequency synthesizer operates at 15.5 GHz over a > 5 percent bandwidth and features a RF enable/disable switch, suitable for On-Off Keying (OOK), Amplitude-Shift Keying (ASK) and Pulse-Position Modulation (PPM) techniques. The switching capability features > 30 dB on/off isolation and < 50 nsec transition time. The THOR-16000 is housed in a hi-rel, rugged package (2.5" × 1.1" × 0.4") of aluminum alloy, with an operating temperature range of -40° to +85°C. Additional characteristics of the unit include low phase noise, low spurs and low power consumption (+5 VDC at < 200 mA).

EM Research Inc.,
Reno, NV (775) 345-2411,
www.emresearch.com.

RS No. 239

Coaxial Resonator Oscillator

MODCO now offers a fundamental coaxial resonator oscillator for 7 GHz applications. The new MCR7200-7350MC tunes 7200 to 7350 MHz. Phase noise is -94 dBc at 10 kHz offset, Vcc is 8.0, Vt is 0.5 to 7.5 and Icc is 21 mA. Industry standard is 0.5 inch package.

Modco Inc.,
Sparks, NV (775) 331-2442,
www.modcoinc.com.

RS No. 240

Low Phase Noise VCO

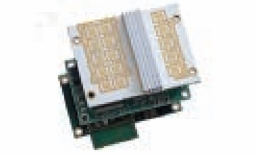
This RoHS compliant voltage-controlled oscillator (VCO) model CRO2690A-LF operates at 2690 MHz with a tuning voltage range of 0.5 to 4.5 VDC. This VCO features a typical phase noise of -114 dBc/Hz at 10 KHz offset and a typical tuning sensitivity of 7 MHz/V. The CRO2690A-LF is designed to deliver a typical output power of 4 dBm at 5 VDC supply while drawing 22 mA (typical) over the temperature range of -40° to 85°C. This VCO features typical second harmonic suppression of -15 dBc and comes in Z-Comm's standard MINI-16-SM package measuring 0.5" × 0.5" × 0.22".

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

RS No. 241

System

24 GHz FMCW Radar



The FMCW radar system operates in the 24 GHz band and can easily be adapted to customer applications, offering efficiency and flexibility in product development. A fully configurable Digital Direct Synthesis (DDS) Controller in combination with a fast locking phase-locked loop (PLL) enables users to define an arbitrary waveform that suits the specific target environment. With an output power of 20 dBm (EIRP) targets up to a distance of 70 m can be detected. The radar system is built out of front-end, baseband and DC-modules that can be replaced depending on the desired functionality.

IMST GmbH,
Kamp-Lintfort, Germany
+49 2842 981 0, www.imst.com.

RS No. 242

Test Equipment

RF Handheld Vector Network and Spectrum Analyzer Tool

VENDORVIEW

Anritsu Co. introduces the MS202xB/MS203xB VNA Master™ series, the highest-performing handheld RF vector network analyzers (VNA) in the industry. With frequency coverage from 500 kHz to 4/6 GHz, and featuring multi-instrument functionality in a rugged, lightweight design that can withstand harsh environments, the MS202xB/MS203xB allow field engineers to find faults, optimize mission-critical systems, and maintain more sites while reducing operating and capital equipment expenditures. VNA Master provides field engineers with a powerful S-parameter measurement tool for on-site installation and maintenance activities, especially two-port transmission measurements such as filter tuning, component test and phase matching cables.

Anritsu Co.,
Richardson, TX (972) 644-1777,
www.anritsu.com.

RS No. 243

MINIATURE FOOTPRINT VCO'S

SYNERGY®
MICROWAVE CORPORATION

Model	Frequency Range (MHz)	Tuning Voltage (VDC)	DC Bias VDC @ I [Typ.]	Phase Noise @ 10 kHz (dBc/Hz) [Typ.]	Size (Inch)
DCO Series					
DCO50100-5	500 - 1000	0.3 - 15	+5 @ 26 mA	-100	0.3 x 0.3 x 0.1
DCO7075-3	700 - 750	0.5 - 3	+3 @ 10 mA	-108	0.3 x 0.3 x 0.1
DCO80100-3	800 - 1000	0 - 3	+3 @ 15 mA	-105	0.3 x 0.3 x 0.1
DCO80100-5	800 - 1000	0.5 - 8	+5 @ 21 mA	-111	0.3 x 0.3 x 0.1
DCO100200-5	1000 - 2000	0.5 - 24	+5 @ 30 mA	-95	0.3 x 0.3 x 0.1
DCO1198-8	1195 - 1205	0.5 - 8	+8 @ 24 mA	-115	0.3 x 0.3 x 0.1
DCO170340-5	1700 - 3400	0.5 - 24	+5 @ 24 mA	-90	0.3 x 0.3 x 0.1
DCO200400-5 DCO200400-3	2000 - 4000	0.5 - 18	+5 @ 35 mA +3 @ 35 mA	-90 -89	0.3 x 0.3 x 0.1
DCO300600-5 DCO300600-3	3000 - 6000	0.5 - 18	+5 @ 35 mA +3 @ 35 mA	-80 -78	0.3 x 0.3 x 0.1
DCO400800-5 DCO400800-3	4000 - 8000	0.5 - 18	+5 @ 35 mA +3 @ 35 mA	-78 -76	0.3 x 0.3 x 0.1
DCO432493-5 DCO432493-3	4325 - 4950	0.5 - 11	+5 @ 17 mA +3 @ 17 mA	-88 -86	0.3 x 0.3 x 0.1
DCO450820-5	4500 - 8200	0.5 - 14	+5 @ 22 mA	-77	0.3 x 0.3 x 0.1
DCO473542-5 DCO473542-3	4730 - 5420	0.5 - 22	+5 @ 20 mA +3 @ 20 mA	-88 -86	0.3 x 0.3 x 0.1
DCO490517-5 DCO490517-3	4900 - 5175	0.5 - 5	+5 @ 22 mA +3 @ 22 mA	-88 -86	0.3 x 0.3 x 0.1
DCO495550-5 DCO495550-3	4950 - 5500	0.5 - 12	+5 @ 22 mA +3 @ 22 mA	-87 -85	0.3 x 0.3 x 0.1
DCO579582-5	5780 - 5880	0.5 - 10	+5 @ 27 mA	-91	0.3 x 0.3 x 0.1
DCO608634-5 DCO608634-3	6080 - 6340	0.5 - 5	+5 @ 22 mA +3 @ 22 mA	-86 -84	0.3 x 0.3 x 0.1
DCO615712-5 DCO615712-3	6150 - 7120	0.5 - 18	+5 @ 22 mA +3 @ 22 mA	-85 -83	0.3 x 0.3 x 0.1

Model	Frequency Range (GHz)	Tuning Voltage (VDC)	DC Bias VDC @ I [Typ.]	Phase Noise @ 10 kHz (dBc/Hz) [Typ.]	Size (Inch)
DXO Series					
DXO810900-5 DXO810900-3	8.1 - 8.925	0.5 - 15	+5 @ 26 mA +3 @ 26 mA	-82 -80	0.3 x 0.3 x 0.1
DXO900965-5 DXO900965-3	9.0 - 9.65	0.5 - 12	+5 @ 22 mA +3 @ 22 mA	-80 -78	0.3 x 0.3 x 0.1
DXO10701095-5	10.70 - 10.95	0.5 - 15	+5 @ 21 mA	-82	0.3 x 0.3 x 0.1
DXO11441200-5	11.44 - 12.0	0.5 - 15	+5 @ 23 mA	-82	0.3 x 0.3 x 0.1
DXO11751220-5	11.75 - 12.2	0.5 - 15	+5 @ 24 mA	-80	0.3 x 0.3 x 0.1

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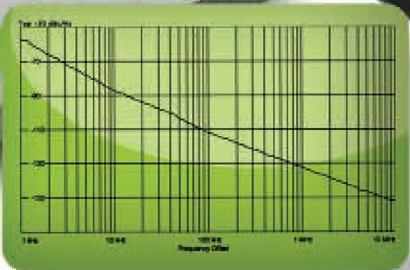

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IMS2011 offers technical paper sessions, interactive forums, plenary and panel sessions, workshops, short courses, industrial exhibits, application seminars, historical exhibits, and a wide array of other technical and social activities including a guest program. The Awards Banquet and Crab Feast are two of the highlights of the social activities. Collocated with IMS2011 are the RFIC Symposium (www.rfic2011.org) and the ARFTG Conference (www.arftg.org), which comprise the Microwave Week 2011 technical program.

PAPER SUBMISSION INSTRUCTIONS: Authors are invited to submit technical papers describing original work on radio-frequency, microwave, and millimeter-wave theory and techniques. The deadline for submission is **December 3, 2010**. **Late papers will not be reviewed.** Please refer to the IMS2011 website (www.ims2011.org) for complete submission information.

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Friday	*Friday*	*Monday*	*Friday*	*Friday*	*Sun. – Fri.*
Proposal Submission Deadline	**Paper Submission Deadline**	**Paper Disposition Notification**	**Manuscript Submission Deadline**	**Notes Submission Deadline**	**Microwave Week**
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4. CAD Algorithms and Techniques
5. Linear Device Modeling
6. Nonlinear Device Modeling
7. Nonlinear Circuit and System Simulation

Passive Components
8. MEMS Components and Technologies
9. Transmission Line Elements
10. Planar Passive Filters and Multiplexers
11. Non-planar Passive Filters and Multiplexers
12. Active, Tunable, and Integrated Filters
13. Ferroelectric, Ferrite, and Acoustic Wave Components
14. MEMS Components and Technologies

Active Components
15. Semiconductor Devices and Monolithic ICs
16. Signal Generation
17. Frequency Conversion and Control
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1. Publication Title: Microwave Journal	2. Publication Number: 3 9 6 - 2 5 0	3. Filing Date: September 30, 2010
4. Issue Frequency: Monthly	5. Number of Issues Published Annually: 12	6. Annual Subscription Price: Domestic: $120.00 Foreign: $200.00

7. Complete Mailing Address of Known Office of Publication *(Not printer) (Street, city, county, state, and ZIP+4®)*
HORIZON HOUSE PUBLICATIONS, INC.
685 Canton Street
Norwood, MA 02062-2610

Contact Person: Michelle Barry
Telephone: 781-619-1930

8. Complete Mailing Address of Headquarters or General Business Office of Publisher *(Not printer)*
same as above

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14. Issue Date for Circulation Data Below: YTD September 30, 2010

15. Extent and Nature of Circulation		Average No. Copies Each Issue During Preceding 12 Months	No. Copies of Single Issue Published Nearest to Filing Date
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b. Paid Circulation *(By Mail and Outside the Mail)*	(1) Mailed Outside-County Paid Subscriptions Stated on PS Form 3541. (Include paid distribution above nominal rate, advertiser's proof copies, and exchange copies)	28,219	29,104
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f. Total Distribution *(Sum of 15c and 15e)*	▶	38,062	41,567
g. Copies not Distributed *(See Instructions to Publishers #4 (page #3))*	▶	65	97
h. Total *(Sum of 15f and g)*	▶	38,127	41,664
i. Percent Paid *(15c divided by 15f times 100)*	▶	94.30%	88.22%

16. Publication of Statement of Ownership
☑ If the publication is a general publication, publication of this statement is required. Will be printed in the November, 2010 issue of this publication. ☐ Publication not required.

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PS Form **3526**, September 2007 *(Page 2 of 3)*

THE BOOK END

Multigigabit Microwave and Millimeter-Wave Wireless Communications

Jonathan Wells

For decades, microwave radios in the 6 to 50 GHz bands have been providing wireless communications. Recently, newer technologies at the 60 to 100 GHz mmWave bands have taken advantage of new wireless regulations that are designed to enable ultra-high capacity communications. *Multigigabit Microwave and Millimeter-Wave Wireless Communications* explores this area in depth and offers the latest details on multigigabit wireless communications. The book places emphasis on practical use and applications, but also provides a thorough explanation of important technological underpinnings to give readers a complete understanding of the subject.

Multigigabit Microwave and Millimeter-Wave Wireless Communications is very current and covers the newer standards such as 60 GHz WLAN, WirelessHD, IEEE 802.15.3c and 802.11ad, ECMA-387/ISO/IEC 13156 and WiGig. It first reviews the high data rate radio spectrum and standards before discussing high-frequency propagation, high data rate wireless systems and multigigabit microwave radios. It then covers more specific types of mmWave radio designs and applications in detail, including future directions and trends.

The 60 GHz band is covered in detail, including frequency bands, channel sizes, propogation, applications, standards and systems. Another chapter covers similar topics for 70/80 GHz radios and higher bands. Finally, high data rate wireless link design is covered, including link budget and fade margin with examples plus environmental effects.

This is a well written book covering a wide range of critical topics, from the high frequency electromagnetic spectrum and high data rate mmWave radios, to wireless link margins and path profiling. It is compact, focused and can be digested in a short period of time if one is already familiar with mmWave technology.

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SPST								
0.2 – 2	SW1-002020RN1NF	1.7	70	1.6:1	10/10	20	35	35/70
2 – 8	SW1-020080RN1NF	2	80	1.7:1	10/10	20	35	35/70
4 – 12	SW1-040120RN1NF	2.2	80	1.7:1	10/10	20	35	35/70
2 – 18	SW1-020180RN1NF	3	80	2:1	10/10	20	35	35/70
1 – 18	SW1-010180RN1NF	3	70	2:1	10/10	20	35	35/70
SP2T								
0.2 – 2	SW2-002020RN1NF	1.5	70	1.6:1	10/10	20	35	60/60
2 – 8	SW2-020080RN1NF	1.8	80	1.7:1	10/10	20	35	60/60
4 – 12	SW2-040120RN1NF	2.2	80	1.7:1	10/10	20	35	60/60
2 – 18	SW2-020180RN1NF	2.8	80	2:1	10/10	20	35	60/60
1 – 18	SW2-010180RN1NF	3	70	2:1	10/10	20	35	60/60
SP3T								
0.2 – 2	SW3-002020RN1NF	1.6	70	1.6:1	20/20	150	180	85/85
2 – 8	SW3-020080RN1NF	1.9	80	1.7:1	20/20	150	180	85/85
4 – 12	SW3-040120RN1NF	2.4	90	1.7:1	20/20	150	180	85/85
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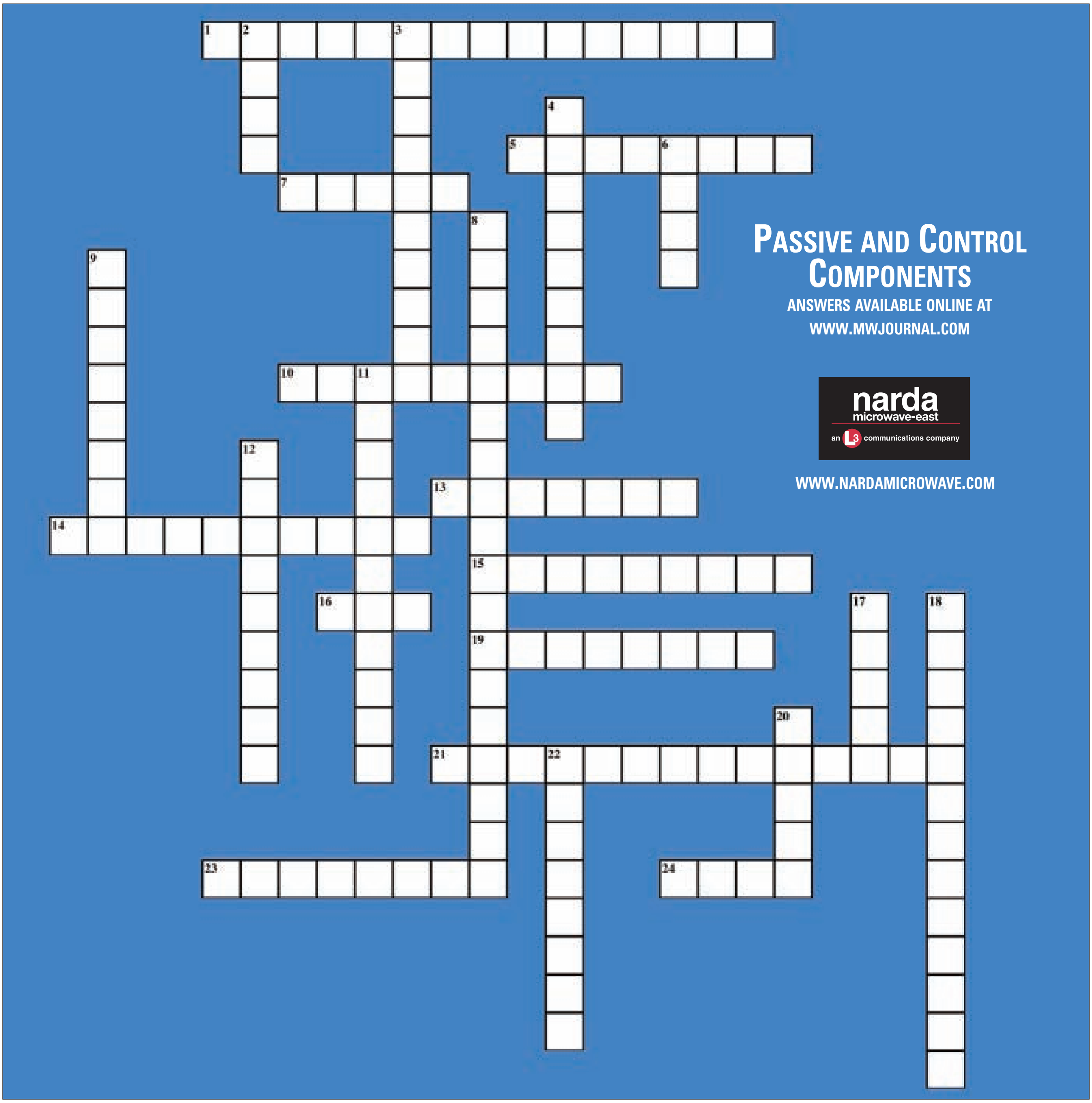

ACROSS

1 GaAs PHEMT FET type that allows for digital logic integration on MMICs (2 words)

5 Currently the most widely used device technology for RF switches (2 words)

7 Type of transformer that can convert signals that are balanced to signals that are unbalanced and vice versa

10 Patented specialty SOI process that uses a very thin layer of Si on a sapphire substrate

13 Passive device used to couple a specific proportion of the power travelling in one transmission line out through another port

14 $=-20\log|\Gamma|$ dB (2 words)

15 Magnitude of a signal that gets coupled across an open circuit

16 Short for silicon-on-sapphire

19 Ron·Coff product is used as a figure of merit for which device type?

21 Passive device that selects only the desired band of frequencies (3 words)

23 Passive device that allows bi-directional communication over a single channel

24 $= \frac{1+|\Gamma|}{1-|\Gamma|}$

DOWN

2 Nano-electromechanical systems

3 Three-port passive device used to control the direction of signal flow in a circuit

4 Passive device made from two metal plates separated by an insulator

6 Micro-electromechanical systems

8 SOI (3 words)

9 Solid-state switch type in use since the 1950s and still very well suited for high power, high frequency applications (2 words)

11 Device that transfers electrical energy from one circuit to another through inductively coupled conductors

12 Newer compound semiconductor switch technology that could compete with PIN diodes for high power applications (2 words)

17 Switch technology that switches states quickly but typically suffers from gate lag

18 Loss of signal power resulting from the insertion of a device in a transmission line (2 words)

20 Nonlinear device that accepts as its input two different frequencies and presents at its output a mixture of signals

22 Passive device that implements frequency domain multiplexing

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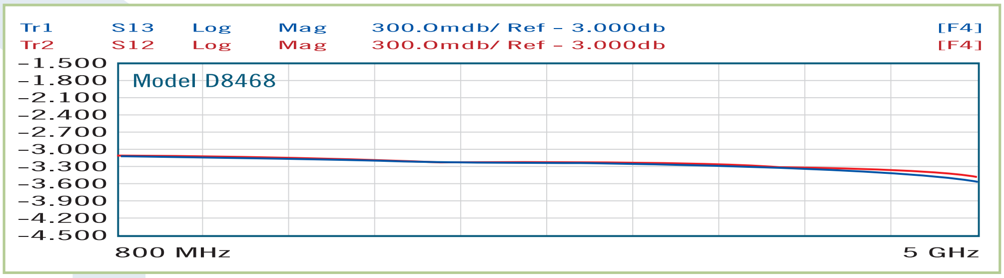
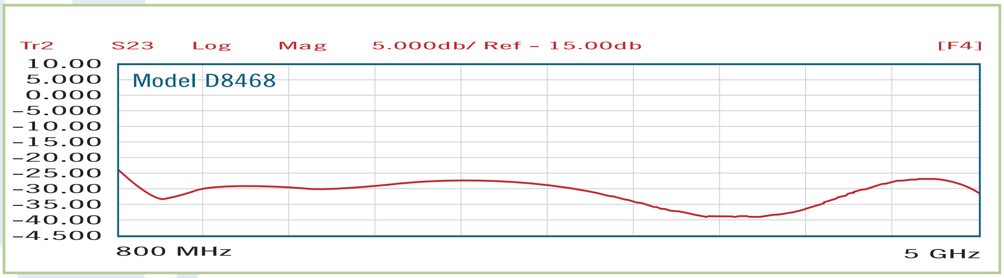


Isolation



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D8611	2-Way	400-3000	200	0.6	1.35:1	15	4.7 x 2.0 x 0.8
D7823	2-Way	500-2500	200	0.4	1.35:1	15	4.7 x 2.0 x 0.8
D8414	2-Way	600-3000	200	0.5	1:35:1	15	4.0 x 1.9 x 1.0
D8378	2-Way	500-2000	800	0.4	1:35:1	15	4.0 x 1.9 x 1.37
D8468	2-Way	800-5000	150	0.6	1:35:1	15	3.4 x 1.4 x 0.87
D8294	2-Way	1000-4000	100	0.5	1:30:1	15	3.7 x 1.9 x 0.87

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RF/Microwave Mobile Communications 2010

From HSPA to LTE and Beyond: Mobile Broadband [R]evolution

by Jean-Pierre Bienaimé
UMTS Forum Chairman

A Special Supplement to

Microwave Journal

Mobile Communications

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Posted under Canadian international publications mail agreement #0738654

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From HSPA to LTE and Beyond: Mobile Broadband Evolution

The UMTS Forum (umts-forum.org) is an industry association committed to helping all participants in the mobile broadband ecosystem understand and profit from the opportunities of 3G/UMTS networks and their Long Term Evolution (LTE). As a Market Representation Partner in the Third Generation Partnership Project (3gpp.org), the UMTS Forum gratefully acknowledges the support of 3GPP in the preparation of this feature.

The promise of 3G mobile broadband has been borne out by more than 630 million UMTS and HSPA subscribers, supported by a vibrant industry ecosystem. Building on these foundations, the first commercial launches of Long Term Evolution (LTE) technology are enhancing the mobile user experience still further. However, even as LTE gains commercial traction, the Third Generation Partnership Project (3GPP) is cementing a new set of standards to shape a new generation of wireless communications.

MOBILE BROADBAND COMES OF AGE

In September this year, a *New York Times* article suggested that the video sharing web site YouTube—owned by search giant Google—generates 160 million mobile views a day. That's almost triple the corresponding figure from a year ago. Mobile now accounts for a significant proportion of all YouTube traffic. Indeed, for many viewers it is their only means of interacting with the site.

Around the same time, the social networking site Facebook claimed more than 150 million active mobile users. What is more, the number of people accessing Twitter from their mobile handset has shot up by 250 percent since the beginning of 2010. And with more than 350,000 fresh users signing up to Twitter every day, 16 percent of newcomers initially access the microblogging service from their phone. For these and millions of other users, the PC is increasingly seen as a secondary means of accessing the Internet.

3G mobile broadband—as delivered by over 350 W-CDMA/HSPA networks globally—has changed the way we work and play. For better or worse, it keeps business travelers connected to the office, wherever they are. It has added an irresistible new dimension to YouTube, Facebook and other services that were originally conceived with fixed Internet users in mind. Furthermore, 3G mobile has enabled the rise of a new breed of on-the-move services that have no counterpart in the fixed world. Since its launch in March 2009, location-based social networking site Foursquare has attracted three million users, with new additions running at nearly 20,000 per day.

Non-voice traffic on 3G wireless networks continues to rise organically, fuelled by a compelling user experience delivered by the current generation of feature-rich mobile devices. Smartphones are everywhere, supported by a fertile development ecosystem. Some 250,000 third-party applications are officially available

Jean-Pierre Bienaimé,
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Just as significant in terms of driving data consumption are PC dongles and netbooks with embedded wireless connectivity, plus a new wave of tablet devices. Aside from its primary nomadic application, Internet connectivity via 3G is increasingly seen as a substitute for fixed mobile broadband in many homes and offices. Indeed, PC usage—from big file transfers to HD-quality video streaming—represents a major chunk of all mobile data consumption.

At the end of September 2010, cellular market information provider Wireless Intelligence confirmed over 630 million subscriptions globally to 3G/UMTS networks. Of these, more than half are customers of high-speed HSPA and HSPA+ networks. Enjoying multi-megabit downlink rates that rival fixed broadband connections at home or at work, many of these customers are voracious data users. They rely on mobile to satisfy their thirst for streaming music and HD video, real-time maps, P2P file exchange and non-stop social networking status updates.

This explosion in mobile broadband usage confirms emphatically the success of 3G. It also underlines the pivotal role of the Third Generation Partnership Project (3GPP)—the body that unites six telecommunications standards bodies from around the world—in continually enhancing the performance of wireless networks to meet evolving market needs. Through successive standards releases, 3GPP offers the wireless industry a coherent framework to serve the evolving needs of its customers while optimising the value of operators' current network investments.

3G/UMTS builds directly on the extraordinary success of the original GSM system that now numbers well over 4.5 billion connections globally. The sheer size and reach of the GSM/UMTS footprint infers obvious economies of scale for both network equipment vendors and terminal equipment manufacturers. This, in turn, defrays development costs. Teamed with global competition, it also realises the possibility of lower end-user pricing for hardware and services, as evidenced by the rise of the $20 handset.

The universal success of GSM also impacts positively on 3G operators and their customers. The close family resemblance between second- and third-generation systems lets subscribers enjoy a transparent experience as their terminal switches seamlessly between 2G and 3G networks according to geographic availability. And for the hundreds of GSM operators—and W-CDMA/HSPA greenfield networks—the business case for evolving to higher data speeds via successive technology iterations is compelling.

Many of the performance enhancements delivered by successive 3GPP releases can be realised by relatively simple, cost-effective software upgrades to operators' existing networks. Even in the case of more fundamental upgrades, this transformation can be achieved while retaining key portions of their network assets, from cell site infrastructure and backhaul to billing and customer care functions.

Faced with the inexorable rise in network traffic, mobile operators must continue to evaluate their options for carrying these massive data volumes more efficiently while catering for further increases in future demand. The needs of today's data-hungry customers are already being met as operators roll out HSPA+ and now LTE, the latest commercial iteration of the 3GPP family that is already live in the US and Europe.

Even as LTE steadily gains commercial traction, 3GPP is currently fine-tuning a new set of standards that will shape a new generation of wireless communications over the next decade and beyond. Here, we briefly review the status of wireless standardisation in 3GPP and corresponding commercial activity. In particular, we examine the objectives and timescales for the ITU's IMT-Advanced project, examining how 3GPP has responded to the ITU's challenge with its own candidate for the 'true' fourth generation of mobile systems.

TOWARDS A NEW GENERATION: LTE-ADVANCED AND 'TRUE' 4G

Standardised in 3GPP Release 8, LTE can be seen as the culmination of a globally co-ordinated development project over the past quarter of a century to create the first truly international broadband multimedia mobile telecommunication system.

Today, the first family of standards derived from the ITU's original IMT concept—IMT-2000, or '3G' as it is commonly known to operators and end-users alike—has delivered voice and broadband data capabilities to almost 800 million subscribers. Of this total, the great majority (over 630 million) are connected to the UMTS/W-CDMA/HSPA family of 3G systems as specified by 3GPP. They are complemented by an estimated 150+ million subscriptions to 1xEV-DO networks, based on the CDMA-based IMT-2000 family member as standardised by 3GPP2.

However, with IMT-2000 requirements now over 10 years old, it is time to set a fresh direction for the next major epoch in mobile communications. User expectations have changed dramatically since the first set of radio interfaces were approved for IMT-2000 in 1999. A decade ago, affordable, ubiquitous, high-speed Internet access—even in the fixed world—was still a dream. Outside the office and academia, access to the web was slow via dial-up connection. On the move, 'mobile data' effectively meant SMS.

> "The requirements for IMT-Advanced are a significant milestone in capability when compared to those of IMT-2000. IMT-Advanced is a leap beyond. It offers new capabilities for the physical layer of the radio interface and brings into play a greater level of radio resource management and control, advanced capabilities from spectrum channel and bandwidth aggregation, and improved performance at all levels, including quality of service aspects. IMT-Advanced represents a wireless telecommunication platform that has the flexibility to accommodate services that are yet to be imagined." (See ***Figure 1***)
>
> Stephen M. Blust, Chairman of ITU-R Working Party 5D

In this context, the ITU's original goals for IMT-2000—data speeds of 384 kbps at pedestrian speeds rising to 2 Mbps indoors—are impressive in themselves. More than a decade on, the ITU's fresh vision for a completely new generation of mobile systems is equally ambitious. A decade from now, the wireless landscape will look very different. Users will access a new breed of ultra-high speed mobile broadband services and applications via a heterogeneous blend of radio ac-

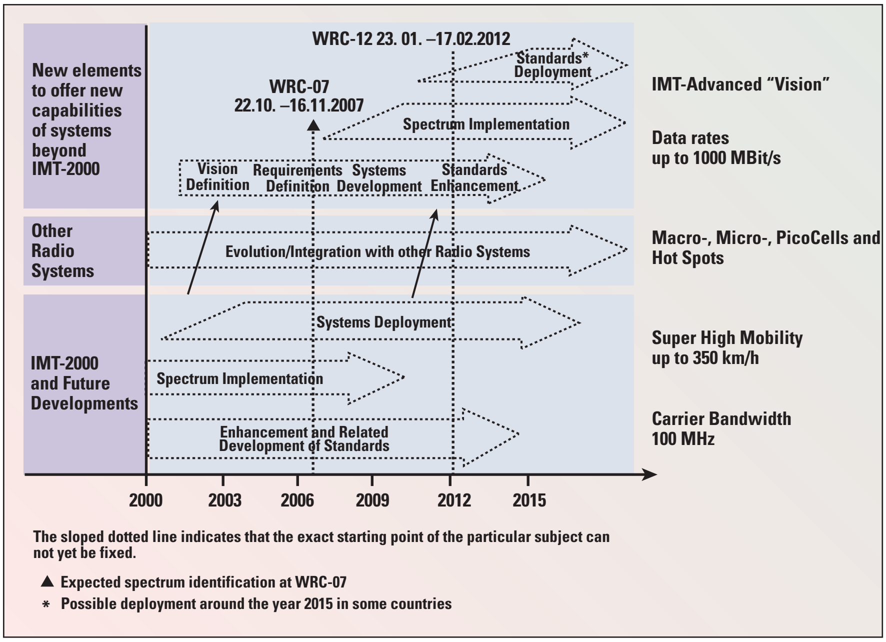

Fig. 1 Phases and expected timeline for IMT-Advanced development and deployment (source: Telefónica 02).

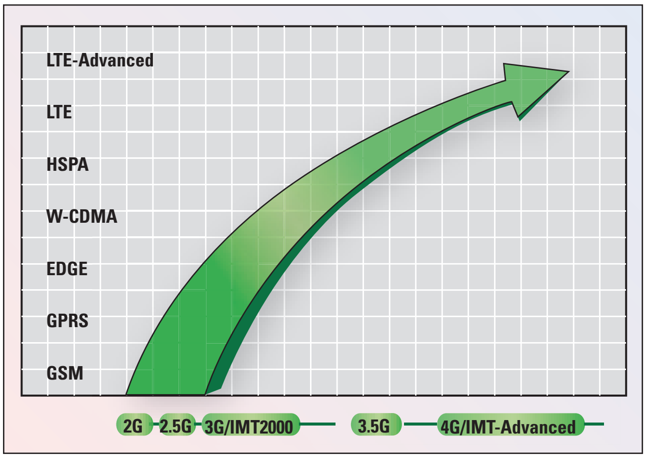

Fig. 2 Meeting the ITU's formal requirements for IMT-Advanced, LTE-Advanced is one of the first true 4G systems (source: 3GPP).

cess methods, network topologies and radical approaches to spectrum usage: a 'network of networks'.

Users will connect and interact with these services via radically new types of terminal devices. Wireless connectivity will not be restricted to phones, PCs and tablets. It will be embedded as a matter of course in domestic appliances, vehicles and consumer electronics devices. Machine-to-machine communications will be ubiquitous, with the 'Internet of things' numbering not millions but billions of end-points.

Building on the extraordinary global success of 3G, including W-CDMA, HSPA and now LTE, the ITU has articulated a new vision for this next era of global wireless communications. Termed 'IMT-Advanced', it is a name that echoes the ambitions of the previous century's original IMT project. This time, however, the goals are still higher.

According to the ITU's definition: "IMT-Advanced is a leap beyond new radio interface physical layer capabilities and brings into play a greater level of radio resource management and control, advanced capabilities for spectrum aggregation, and improved performance at all levels including QoS [Quality of Service] aspects."

"International Mobile Telecommunications-Advanced (IMT-Advanced) systems are mobile systems that include the new capabilities of IMT that go beyond those of IMT-2000," states a contribution from ITU's Radiocommunication Sector (ITU-R) Working Party 5D in March 2008. "Such systems provide access to a wide range of telecommunication services, including advanced mobile services, supported by mobile and fixed networks, which are increasingly packet-based."

The description hints strongly at a heterogeneous future, where mobility is characterised by a total user experience across a mesh of fixed and mobile networks: "IMT-Advanced systems support low to high mobility applications and a wide range of data rates in accordance with user and service demands in multiple user environments. IMT-Advanced also has capabilities for high quality multimedia applications within a wide range of services and platforms, providing a significant improvement in performance and quality of service."

In its earliest discussions about IMT that date back to the early years of this millennium, the ITU was circumspect in describing this new generation of advanced systems as '4G'. By October 2009, however, ITU explicitly used the term when it announced that six candidate technology submissions for 4G had been received in response to an open invitation in March 2008 (see ***Figure 2***).

Currently, several operators and vendors have branded their LTE (and WiMAX) offerings as '4G'. However, this nomenclature is potentially confusing when seen in the light of the ITU's official terminology. LTE and WiMAX certainly offer a significant step forward in terms of data rates compared with previous approaches. Their capabilities, however, fall some way short of the ITU's formal requirements for IMT-Advanced.

As such, LTE and other systems are better described as '3.9G'—the last evolutionary step before LTE-Advanced. From the customer's perspective, discussions about what properly constitutes 3G or 4G are of little interest. The immense, globally coordinated standardisation effort by 3GPP is easily obscured by the seamless simplicity of today's mobile user experience. Whether I am updating my Facebook status or uploading some slides before a meeting, all that I am aware of is an experience that is slicker and quicker than it was a year ago… and the year before that.

A BREAK WITH THE PAST

The 'LTE-Advanced' name reflects its roots in LTE and the systems that preceded it—from HSPA+


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TABLE I

2G/3G/4G TECHNOLOGIES COMPARED

	Technology	Carrier BW	UL Peak Data Rate	DL Peak Data Rate	Latency	Spectrum (MHz)	Peak Spectral Eff. (Bit/s/Hz)
2G	GSM/GPRS EDGE (MCS-9)	200 kHz	56 Kbps 118 Kbps	114 Kbps 236 Kbps	500 ms 300 ms	900/1800	0.17 0.33 EDGE
	W-CDMA	5 MHz	384 Kbps	384 Kbps (2 Mbps)	250 ms	900/1800/ 2100/2600	0.51
	HSPA	5 MHz	5.7 Mbps	14 Mbps	-70 ms	DD/900/ 2100/2600	2.88
3G	HSPA+ (16 QAM) (64 QAM +Dual)	5 MHz	11.5 Mbps	-28 Mbps (42 Mbps)	-30 ms	DD/900/ 2100/2600	12.5
	LTE (Rel.8) (2×2 MIMO)	var. up to 20 MHz	-75 Mbps	-150 Mbps (@ 20 MHz)	-10 ms	DD/900/1800 2100/2600	16.32
	WiMAX IEEE 802.16e	10 MHz	70 Mbps	70 Mbps 134 Mbps	-50 ms	2600/3500	3.7
4G	LTE-Advanced*	var. up to 100 MHz	>500 Mbps	>1 Gbps	<5 ms	IMT	DL: >30 UL: >15
	"IMT-Advanced"	var. up to 100 MHz	270 Mbps 675 Mbps	600 Mbps 1.5 Gbps	<10 ms	IMT	DL: >15 UL: >6.75

Source: 3GPP/Telefonica 02
**To be confirmed with 3GPP Ref. 10 by end 2010*

and HSPA though W-CDMA right back to GSM.

Specified in 3GPP Release 10, LTE-Advanced has been explicitly dimensioned to meet—or even exceed—the requirements of IMT-Advanced. Building on the pyramid of previous releases, LTE-Advanced is also backward compatible with Release 8 (LTE), helping operators to effectively leverage their current wireless investments.

The advanced performance requirements of 4G mandate a break with the past. While interworking with 2G and 3G legacy systems will be supported, 4G demands radically new transmission technologies, plus fresh approaches to spectrum usage. Building on the agenda now being set by LTE, 4G will represent a total break from a circuit-switched world. (By comparison, while a number of approaches are currently being considered, it is likely that at least some operators will initially support voice in LTE via circuit-switched fall-back to 2G/3G.)

So what exactly will 3GPP Release 10 offer? As you would expect, the key dimensions where LTE-Advanced scores over previous technologies include speed, spectral efficiency and flexibility, capacity, coverage and interworking.

Captured in Report ITU-R M.2134, the basic requirements of IMT-Advanced are as follows:

- A high degree of commonality of functionality worldwide while retaining the flexibility to support a wide range of services and applications in a cost efficient manner;
- Compatibility of services within IMT and with fixed networks;
- Capability of interworking with other radio access systems;
- High-quality mobile services;
- User equipment suitable for worldwide use;
- User-friendly applications, services and equipment;
- Worldwide roaming capability;

One of the standout attractions of LTE-Advanced is its ability to exploit variable, ultra wide carrier bandwidths of 40 MHz, right up to 100 MHz. This, in turn, supports extremely high data rates in the mobile environment. A new transmission scheme—based on OFDMA/SC-FDMA plus sophisticated MIMO techniques and other measures—will together achieve theoretical peak data rates of 100 Mbps in high mobility situations.

In stationary environments, this theoretical performance rises as high as 1 Gbps—an order of magnitude faster than LTE. Uplink transmission is similarly enhanced, targeting data rates up to 500 Mbps. To realise these performance targets, Release 10 boosts peak spectral efficiency in both uplink and downlink by an order of magnitude compared with HSPA. Other measures, like relay functionality, will boost cell edge coverage, improving the end-user experience in rural and non-optimal coverage areas.

Physics dictates that data rates of the order of 1 Gbps in 4G systems require bandwidths approaching 100 MHz. This infers that spectrum sharing—either through regulatory measures or technology developments involving spectrally agile systems—will inevitably be part of future considerations about maintaining a competitive environment in mobile broadband.

Of equal appeal to real-world users, LTE-Advanced cuts round-trip latency times to 10 ms or less, a fraction of even the 30 ms (approx.) attained with HSPA+. For comparison, HSPA (corresponding to Releases 5 and 6) manages a latency performance no better than 70 ms, while the original W-CDMA standard (Release 99) lags

[Continued on pg. 26]


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How Mobile Device Users are Impacting the Future of RF Front Ends

In the last decade, cellular mobile devices have undergone dramatic changes. What began as a mobile phone simply used for people to talk or text with one another has now turned into a handheld device that provides multi-functionality such as a phone, web browser, text messenger, camera, gaming unit, MP3 player, and many other useful functions to satisfy our need for information on-the-go. Not only do today's mobile device users want all of these features included, they also want them readily available to them at all times, irrespective of time or location. This type of on-demand mobile technology requires compatibility of multiple frequency bands and modulation standards. This type of complex functionality, along with the consumer's desire for smaller form factors, has placed great demands on mobile designers to deliver products at a lower bill of materials (BOM) cost and within record-breaking time to satisfy the market's expectations generation over generation. Such stringent requirements have forced designers to undergo a change in the way that RF front ends are benchmarked. This article discusses some of these impacts and how a new approach can be embraced to enhance the consumer experience when using a feature-rich mobile device.

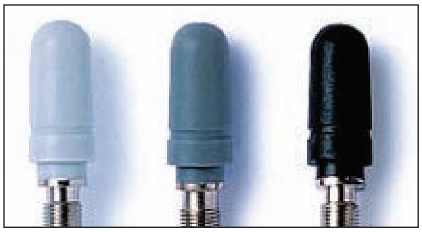


Fig. 1 External stub antennas.

Readers who were in the industry several years ago can remember that voice was the primary driver for performance, and the most widely used modulation format was GSM/GPRS. Handset designs were much larger, with more printed circuit board (PCB) real estate dedicated to the RF front-end section, and performance was the primary focus of the project. The antennas were external to the handset, as depicted in ***Figure 1***, taking the form of a stub or slider that pulled out and retracted with efficiencies that were much better than what a user can find in a handset today. Phones were designed to be operated in voice-only calls with the handset or mobile device held in a fairly predictable position relative to the user's head. The consistency allowed the antenna to be designed in a fairly known environment that allowed for optimization of the design. This is still critical today since the power amplifier (PA) can be a significant drain on talk time, which is directly correlated to a user's experience with a certain model of device as well as the company brand of mobile device. If the designer can optimize the current consumption in a real-world environment, the better positioned the mobile device is in the consumer market. The consistency of antennas and their real-world behavior has given handset designers the flexibility to optimize their design by impedance matching

Jackie Johnson
RFMD, Greensboro, NC

Fig. 2 *PIFA antenna.*

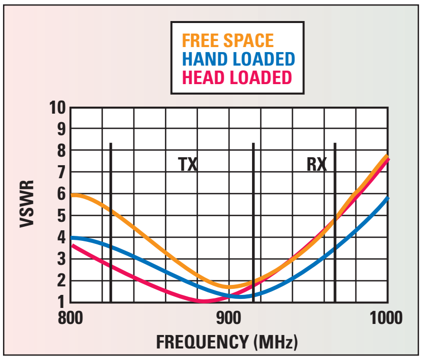


Fig. 3 *VSWR performance of a PIFA antenna in a mobile device.*

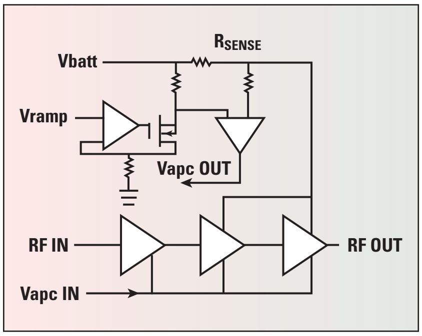


Fig. 4 *Current control block diagram.*

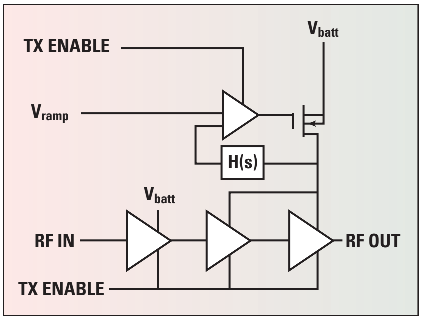


Fig. 5 *Voltage control block diagram.*

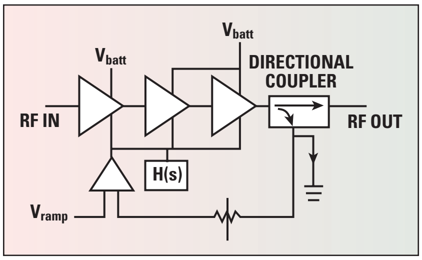


Fig. 6 *Power detection block diagram.*

antennas and PAs in order to deliver the maximum amount of power as efficiently as possible.

MOBILE DEVICES: THAT WAS THEN, THIS IS NOW

Fast forward a couple of years and the mobile device market has changed dramatically. Real estate is now dedicated to applications processors and components that are more focused on software applications than enhancing the consumer experience. Mobile devices are now designed with much smaller form factors and performance has been traded off in many cases to achieve these unique form factors. Handsets today have integrated patch or Planar Inverted F Antennas (PIFA) (illustrated in ***Figure 2***) that in many cases are far less efficient to their predecessors. Even today, some handsets are now reverting back to a stub antenna due to the issues that many designers are facing. This performance versus form factor tradeoff directly impacts a consumer experience in the form of battery life, talk time and network availability since the antenna choices and their environment affect the PA.

One example of how this affects PA performance is in the form of voltage standing wave ratio, or VSWR. Mobile devices today are operated in three basic configurations. Users talk on the mobile device with the handset next to the head in a conventional manner, out in front of their head using the speaker phone and in a free space environment where the phone is not held. These are just three scenarios in which the VSWR response of the antenna can change. In reality, there are numerous configurations determined by the position of the fingers and hand, but for simplicity this article is going to focus only on these three scenarios. The difference in performance of the antenna VSWR is illustrated in ***Figure 3***.

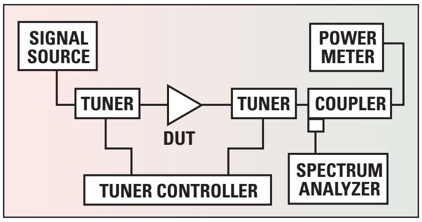
These frequency responses illustrate the different VSWR requirements that the PA faces in a current generation handset. At the band edges, the PA will be exposed to VSWR ranges from 5:1 to 2:1 in this particular mobile device. The RX sensitivity is also affected due to the VSWR performance as well. Common practice used today by many handset designers for benchmarking RF front ends is measuring performance in a 50 ohm lab environment. This method is no longer practical in today's designs due to the unpredictability of the impedances that are seen by the PA. A designer who wants to optimize his or her solution to provide the best talk time to the end user must begin to examine their RF front ends under VSWR conditions.

Standard governance boards such as 3GPP set the requirements for over-the-air (OTA) requirements. These requirements are usually much more relaxed than typical carrier requirements as they require more stringent OTA performance. A typical value a carrier may set for their mobile devices is -11 dB from conducted RF output power. In terms of the GSM 850 standard, this would equal a value of 22 dBm OTA requirement since conducted output power requirement is set to 33 dBm with -11 dB of loss due to antenna efficiency and propagation effects which are frequency dependent. These OTA requirements are directly applicable to the 50 ohm power if RF front ends are benchmarked and compared with these requirements in mind.

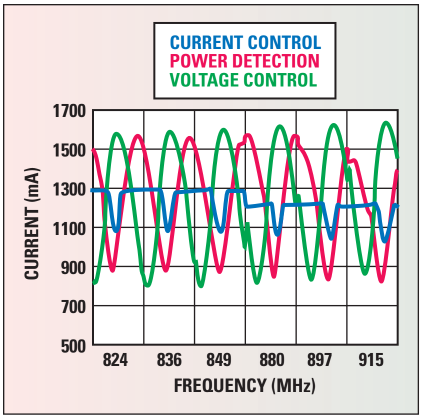
GSM POWER CONTROL ARCHITECTURES' EFFECTS ON TALK TIME

The three different architectures most commonly used in the industry today for GSM mobile applications are current control, voltage control and power detection. A simplified block diagram of each of the three architectures is illustrated in ***Figures 4, 5*** and ***6***.

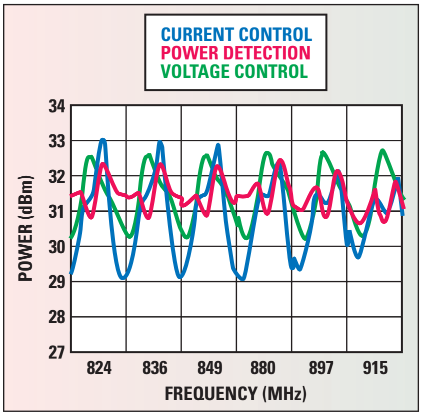
Numerous articles have been published on the theory of these three architectures; therefore, this article will provide only a brief description as background. The current control architecture in Figure 4 is an indirect control scheme in which the current is monitored and held constant. This method relates current to power and provides a very good method of power control as long as the relation between current and power remains constant (which occurs only if the resistance of the load does not change). Power is controlled by adjusting the base bias of the amplifier controlling the gain, which results in power control. Figure 5 is an illustration of voltage control, which is very similar to current control in the sense that it is an indirect method relating voltage to power instead of current. This method—much like



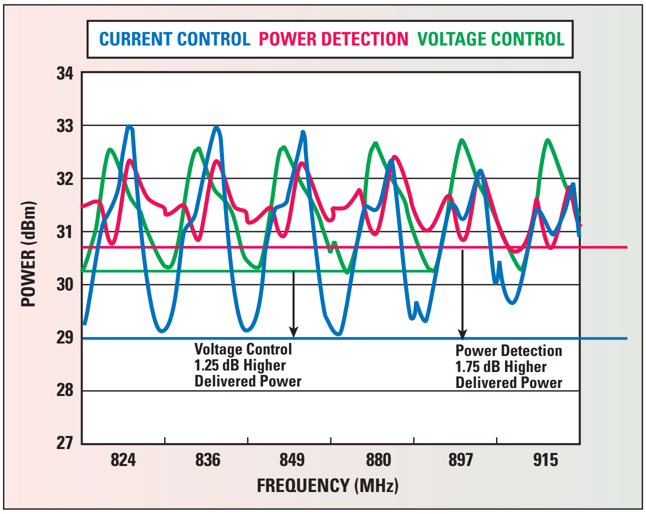
▲ *Fig. 7 Load pull setup.*



▲ *Fig. 8 Current measurement into 3:1 VSWR with 50 Ω power set to 33 dB.*



▲ *Fig. 9 Delivered power into 3:1 VSWR with 50 Ω power set up to 33 dB.*



▲ *Fig. 10 Deltas in delivered power.*

current control—works well as long as the resistance of the load remains constant and the relationship between voltage and power is maintained. The collector voltage is adjusted to control power instead of the base bias like in current control. The final architecture to be compared in this article is power detection, which is illustrated in Figure 6. In this method, power is detected by coupling a portion of the signal back to a detector that compares the output voltage and the reference voltage. The accuracy of this power control scheme is great as well and the mismatch performance is greatly dependent on the directivity of the coupler and the error in the feedback loop. The disadvantage of this architecture, however, is the added output loss of the coupler and the cost of the component, as it takes more circuitry to accomplish the power control function.

After reviewing a very simplistic view of basic power control architectures, one can focus on benchmarking the devices in such a way that reflects real-world performance and has a direct impact on the consumer satisfaction with talk time, battery life and call reception. First, to understand the real-world environment, the antenna performance must be characterized, as illustrated in Figure 3. As mentioned previously, the VSWR can range from 2:1 to 5:1, depending on the end user and how the phone is positioned. Based on these measurements, a good benchmark for comparison is determined to be a 3:1 VSWR. A VSWR of 3:1 is chosen as this will provide a good indication of part performance in the real world without providing unrealistic reflections back to the PA that may distort the results of the comparison. In order to characterize these products appropriately, a load pull must be performed where the designer has precise control over mismatch, phase angle and delivered power accuracy. Such a method is shown in ***Figure 7***.

After examining ***Figures 8*** and ***9***, one can see that even though the power control function is achieved with different architectures, the performance in a real-world environment is very different. What does this mean and why is this so important? First of all, as previously discussed, what really matters is OTA performance, which is a direct correlation with delivered power. As Figure 9 shows, current control is the least desirable solution of the three for maintaining a constant delivered power into the load. There is almost 1.5 dB of difference between current control and power detection in the GSM 850 band. The disadvantage of the power detection scheme is that the current is allowed to increase, whereas the current is maintained reasonably well in the other solution. Although this would result in better talk time in this condition, in a real-world environment it would not.

For example, if the mobile device is operated at 29 dBm, which is the power level that has the greatest probability in a GSM system, the base station would actually request that the handset increase its power level by moving from 29 to 31 dBm since the delivered power cannot be met at current power control level (PCL). This in turn would increase the current consumption and therefore decrease the talk time. Another aspect to consider is the current consumption advantages that can be realized. In a mobile device, if the current control scheme is providing enough delivered output power under these conditions to meet the carrier's OTA requirements, then there is no need to be concerned about power into VSWR. Since the reduction of delivered power is good enough, substantial current consumption savings can be realized with a solution that provides better VSWR performance. In examining ***Figure 10***, consider the following question: If the delivered power could be made equal between all solutions, what effect would this have on the end user?

For voltage control and power detection, the 50 ohm calibration could be set 1 dB lower in power and still meet the equivalent delivered power. ETSI conducted specification specifies that for PCL 5 the power is 33 dBm ±2 dB under nominal conditions. This means that to meet conducted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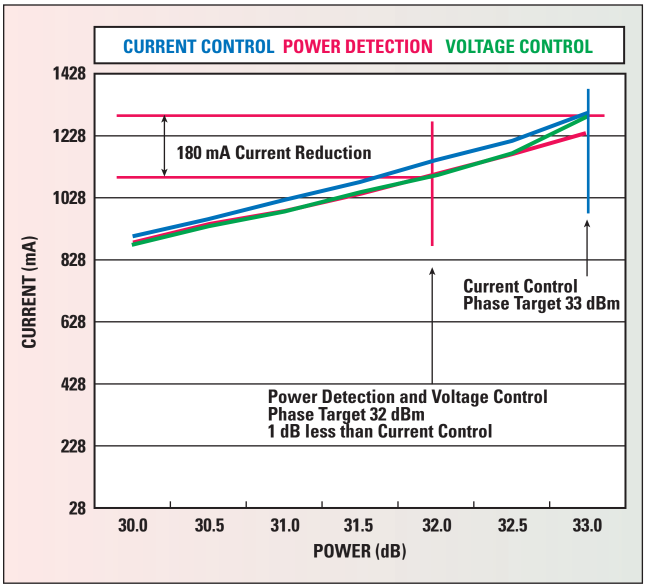
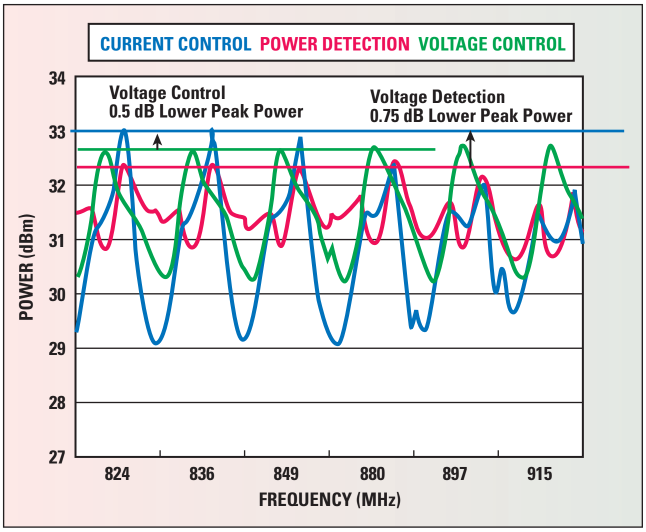


▲ *Fig. 11 50 Ω phasing target delta in current for equivalent delivered output power.*



▲ *Fig. 12 Delta in peak power.*

performance the mobile device can output a minimum of 31 dBm for PCL 5. Understanding the need for margin, the safest level that the mobile device should be calibrated to is 31.5 dBm. If more margin is desired, the designer could gain substantial current savings by phasing the mobile device to 32 dBm in a 50 ohm environment. How this correlates to 50 ohm performance is shown in ***Figure 11***.

In Figure 11 the currents of the three solutions are compared versus output powers. This demonstrates that if the designer can achieve equal delivered output power to meet OTA requirements, then the current control solution would need to be calibrated for 33 dBm of output power compared to 1 dB less for power detection. This would result in 180 mA savings in the 50 ohm environment at full power, which can extend battery life and talk time. This current savings is realized without sacrificing any real-world delivered output power OTA performance. The other advantage to lower phasing targets is more margins to specific absorption rates (SAR) as well as lower harmonic generation since harmonic energy is much lower at 1 dB back-off from full power. This results in fewer emissions issues and faster time to market.

If the designer is not interested in this approach and would like to have more output power, this can also be accomplished with a better VSWR tolerant device. The concern with higher output powers that every designer shares is the possibility of failing the SARs requirement for radiating energy in a multi-slot GPRS case. A much better, well-behaved VSWR tolerant device allows the handset to operate at higher power levels while still meeting SAR requirements under these conditions by limiting the amount of output power that is delivered in a low impedance state (see ***Figure 12***).

Figure 12 illustrates that if a mobile device designer would like to optimize OTA performance, the phasing target could be increased in the mobile device 0.5 to 0.75 dB higher in the voltage and power detection solution compared to the current control. Phasing for higher targets statistically compromise SAR performance. As Figure 12 shows, however, the peak power swings are now equal for all three solutions, and the 50 ohm set power is much higher than the current control solution. This allows the designer to develop a superior product compared to the competition when compared at the carrier for OTA capabilities.

The final consideration is the tradeoff between transmit (TX) and receive (RX) performance and the ability to customize performance based on region. From Figure 3, the VSWR plot of a mobile antenna, there is room to shift the tuning to trade off TX performance for RX performance, if desired. Examining the purple trace, in the case where the phone is head loaded, one can see that if the GSM 850 TX and RX performance was degraded slightly by shifting the frequency response higher in frequencies, then the GSM 900 RX VSWR would actually improve. Having a VSWR tolerant TX path can allow the designer to have the flexibility to make the tradeoffs between the parameters that are most important in his or her particular design.

In conclusion, the importance of benchmarking solutions under mismatch conditions needs to be seriously considered. This method opens a new way of thinking that exposes designers to tradeoffs that can be made at the system level that may have not been considered before. Only examining solutions based on 50 ohm lab testing can and will cause misconceptions on selecting the right architecture for one's design. In Figure 10, it is clear that almost all three solutions can perform the function and are very similar in performance. Although this is true in the 50 ohm environment, it is not the case in real world application. Considering OTA performance can result in greater flexibility for the designer to customize their product for better current consumption, high OTA power, or RX performance. All of these options are revealed if the designer is open to a new way of benchmarking RF front ends and making decisions that truly affect consumer satisfaction in terms of fewer dropped calls and longer battery life. As the end consumer experience improves, so does the brand image of that particular mobile device, which results in greater demand from consumers and higher adoption rates by the carriers. ■

A High Linearity Darlington Intermediate Frequency (IF) Amplifier for Wide Bandwidth Applications

This article describes the design of a very high linearity, wideband intermediate frequency (IF) amplifier. The design has a very flat gain response and most of the circuit components are integrated. It achieves a third-order intercept point (IP3) greater than 43 dBm and a noise figure (NF) less than 2 dB across a wide frequency range of 30 to 1000 MHz.

IF amplifiers are used in many applications, including base stations, cable TV and instrumentation. High linearity is one of the most desired features of an IF amplifier. New applications, such as amplifiers used in cable TV, also require very flat gain and low noise. Darlington amplifiers with single or dual RF feedback topologies have been shown to have higher gain, flatness and linearity over a wide bandwidth.[1,2] Various device process technologies, including HBT, SiGe and PHEMT among others, have been used in the past. Each technology has its unique advantages. Different active, dynamic and passive biasing techniques have also been used to improve performance over temperature and over the frequency band of operation.[3,4] Capacitive peaking techniques are used to further increase the bandwidth, with a trade off in input and output return loss.[5] The drawback in most high linearity designs is the big trade off in noise figure (NF).

The design of a high linearity Darlington RF feedback amplifier with less than 2 dB NF has been achieved. The design uses a source capacitive peaking technique for optimum gain flatness across a wide band. Source degeneration inductors are used for improving the input and output return loss and stability at lower frequencies. Additional stability improvement circuits are used to ensure unconditional stability at higher frequencies.

CIRCUIT DESIGN

The frequency range of operation is from 30 to 1000 MHz. With a Darlington design topology, the first and second stages can be biased at different conditions. Individual voltage and current adjustment of each stage provides extra flexibility for performance optimization. The Darlington configuration provides twice the gain bandwidth product over single stage circuit topologies. Furthermore, good input and output return loss across a wide frequency range can be achieved with RF feedback optimization.

The IP3 can be further improved by presenting different impedance terminations to the device.[6] When more voltage and current are available, the drain voltage and current can be increased to improve IP3 and the 1 dB compression point (P_{1dB}) with minimal impact on other performance parameters.

The device process used in this design is a depletion (D)-mode low noise pseudomorphic high electron mobility transistor (PHEMT) with a thin film resistor (TFR). The PHEMT process inherently has low noise and very high linearity. This makes it suitable for this specific application. However, the design shown in this article is not limited to this type of process.

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Figure 1 shows the circuit schematic of the die. Resistors at the sources (R_4 and R_6) are used to set the positive source DC voltage and the drain currents. The value of these resistors can be easily calculated using the current equation

$$I_{DS} = I_{DSS}\left(1 - \frac{V_{GS}}{V_p}\right)^2 \qquad (1)$$

where the gate-to-source voltage (V_{GS}) and the pinch-off voltage (V_p) values are both negative. Once the drain bias current (I_{DS}) is determined, V_{GS} can be calculated. The gate voltage is chosen to be small. It is applied through the resistive divider (R_1 and R_2) and a very large value of shunt resistor (R_3) connecting to the gate. Given the gate voltage, the source voltage and resistor can be calculated. The total current consumption is 100 mA at 5 V supply voltage. It is also important to consider the thermal design of the output transistor. For this reason the LNFET device layout is optimized using thermal calculators.

Two of the design constraints, NF and input return loss, are the main factors in determining the first stage design. An optimum device sizing and inductive degeneration technique are used for simultaneously optimizing noise figure and input match. The second stage is designed for optimum IP3 performance. The biasing conditions for each stage are crucial in determining the best overall IP3. One can choose the optimum biasing conditions such that second-order and third-order intermodulation products are cancelled.[6]

The drain current is controlled by the gate and drain voltages. In the small-signal case, it can be represented by a two-dimensional Taylor series expansion as an incremental drain current[7,8]

$$\begin{aligned} i_d(v_g, v_d) &= I_d(v_g, v_d) - I_d(v_{g0}, v_{d0}) \\ &= \frac{\partial I_d}{\partial V_g} v_g + \frac{\partial I_d}{\partial V_g} v_d + \frac{1}{2}\left[\frac{\partial^2 I_d}{\partial V_g^{\,2}} v_g^{\,2} + 2\frac{\partial^2 I_d}{\partial V_g \partial V_d} v_g v_d + \frac{\partial^2 I_d}{\partial V_d^{\,2}} v_d^{\,2}\right] \\ &+ \frac{1}{6}\left[\frac{\partial^3 I_d}{\partial V_g^{\,3}} v_g^{\,3} + 3\frac{\partial^3 I_d}{\partial V_g \partial V_d^{\,2}} v_g v_d^{\,2} + 3\frac{\partial^3 I_d}{\partial V_g^{\,2} \partial V_d} v_g^{\,2} v_d + \frac{\partial^3 I_d}{\partial V_d^{\,3}} v_d^{\,3} + \cdots \right] \end{aligned} \qquad (2)$$

where V_{g0} and V_{d0} are the DC bias voltages and v_g and v_d are the small-signal gate and drain voltages.

If the higher order terms are ignored and the coefficients simplified, the small-signal incremental drain current can be written as

$$\begin{aligned} i_d(v_g, v_d) &= g_m v_g + G_{ds} v_d + \frac{1}{2} g'_m v_g^{\,2} + \frac{1}{2} G'_{ds} v_d^{\,2} \\ &+ m_{11} v_g v_d + \frac{1}{6} g''_m v_g^{\,3} + \frac{1}{6} G''_{ds} v_{ds}^{\,3} \\ &+ \frac{1}{2} m_{12} v_g v_d^{\,2} + \frac{1}{2} m_{12} v_g^{\,2} v_d \end{aligned} \qquad (3)$$

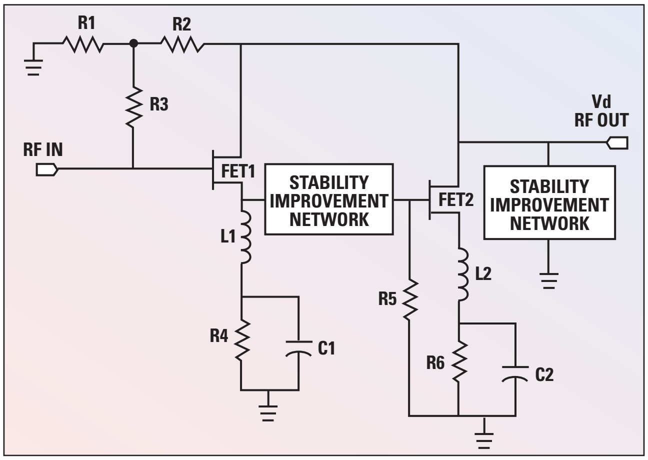

Fig. 1 Die circuit schematic.

where the coefficients g_m, g'_m and g''_m are the transconductance and its first and second derivatives with respect to v_g.

$$g_m = \frac{\partial I_d}{\partial V_g},\ g'_m = \frac{\partial^2 I_d}{\partial V_g^{\,2}} \text{ and } g'' = \frac{\partial^3 I_d}{\partial V_g^{\,3}} \qquad (4)$$

G_d, G'_d and G''_d are the drain-to-source transconductance and its first and second derivatives with respect to v_{ds}.

$$G_d = \frac{\partial I_d}{\partial V_d},\ G'_d = \frac{\partial^2 I_d}{\partial V_d^{\,2}} \text{ and } G''_d = \frac{\partial^3 I_d}{\partial V_d^{\,3}} \qquad (5)$$

And m_{11}, m_{12} and m_{21} are cross terms defined as

$$m_{11} = \frac{\partial^2 I_d}{\partial V_g \partial V_d},\ m_{12} = \frac{\partial^3 I_d}{\partial V_g \partial V_d^{\,2}} \text{ and } m_{21} = \frac{\partial^3 I_d}{\partial V_g^{\,2} \partial V_d} \qquad (6)$$

The drain to source transconductance, in the saturation region where V_{ds} is high, can be assumed to be small. The cross terms above are also generally small and can be ignored. Therefore, the id equation above can be further simplified to

$$i_d(v_g) = g_m v_g + \frac{1}{2} g'_m v_g^{\,2} + \frac{1}{6} g''_m v_g^{\,3} \qquad (7)$$

The above equation gives rough guidance for optimizing the second-order intercept point (IP2) and IP3. It shows that the lowest second-order and third-order distortion products are achieved when the first and second derivatives of transconductance, g'_m and g''_m, are minimized. It is ideal to have both IP2 and IP3 products lowered. This can be accomplished by carefully selecting the bias points as well as optimizing the amplifier transconductance profile over a wide bias range. The analysis above also shows the sensitivity of this method to wafer uniformity and gain profile. The sensitivity can be reduced by process parameters[9] as well as topology selection and feedback techniques.[2]

The Darlington topology with source degeneration allows gain optimization and high linearity. Capacitors C1 and C2 on the die are chosen for achieving the best gain flatness without seriously degrading input and output return losses. These capacitors are also used to bypass the source resistors and hence improve the NF. R1 and R2 on the die are used for setting the gate voltages.

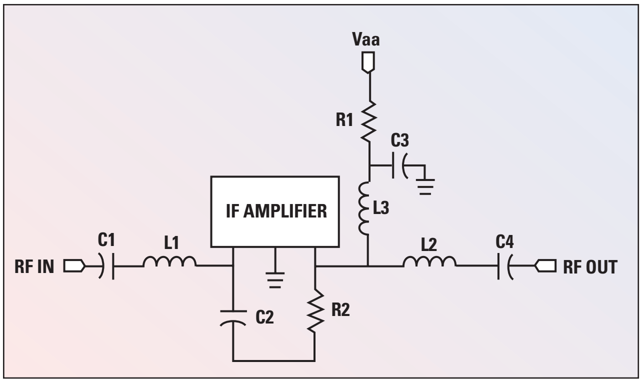

Fig. 2 Evaluation board schematic.

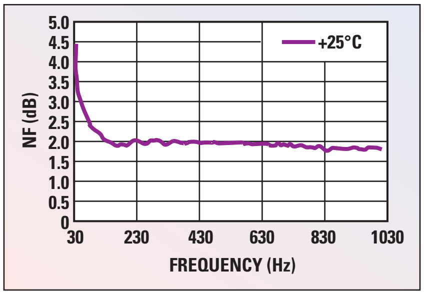

Fig. 3 Noise figure vs. frequency.

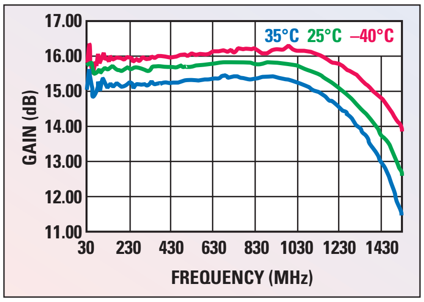

Fig. 4 Small-signal gain vs. frequency.

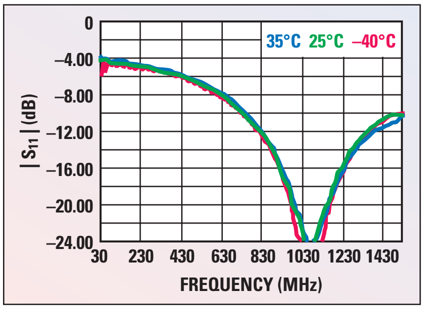

Fig. 5 Input return loss.

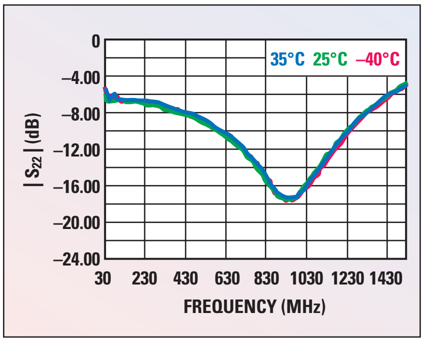

Fig. 6 Output return loss.

Figure 2 shows the test board schematic with external components. The external RF feedback resistor, R2, is chosen for the best input and output return losses and gain trade-off. C1, C2 and C4 are used for DC blocking.

DESIGN FOR STABILITY

Typical specifications dictate unconditionally stable operation up to 18 GHz. This amplifier is designed for unconditionally stable operation, including the external components and biasing under all conditions. For this purpose, various stability design techniques have been employed and integrated into the amplifier. In order to solve stability problems at low to operating frequencies, a source inductor of some value is often used.

COMPONENT SELECTION CONSIDERATIONS

External feedback circuit components R2 and C2 can be tuned if a gain adjustment is needed. The input and output matching networks are composed of L1 and L2, respectively. The input and output matching circuits are designed to be centered at approximately 700 to 1200 MHz. This can also be tuned by the external matching components L1, C1 and L2, C4.

MEASUREMENT RESULTS

An amplifier was fabricated and tested on the test board shown previously with a supply voltage V_{dd} of 5 V and a total supply current I_d = 100 mA. ***Figure 3*** shows the measured noise figure in a bandwidth of 30 to 1030 MHz. The input connector and board trace are not de-embedded from the measurement. The loss of the input transmission line was measured as 0.1 dB in this band. The NF is measured as approximately 2 dB from 100 to 1000 MHz, including the input transmission line loss. ***Figure 4*** shows the measured amplifier gain as a function of frequency at several temperatures with an input power of -20 dBm. The gain is measured to be 15.6 dB at 450 MHz. The plots show that the gain changes less than 1 dB across the whole temperature range of -40° to 85°C.

Figures 5 and ***6*** show the measured input and output return losses, respectively, with an input power of -20 dBm. The output third-order intercept point (OIP3) measured within the operating band was above 43 dBm. The OIP3 measurements were done using two signal sources with frequencies of 433.25 and 449.25 MHz at P_{out} = 5 dBm per tone.

CONCLUSION

Many performance aspects must be considered in the design of IF amplifiers. The next generation IF amplifiers require careful topology selection and competitive design techniques to meet the increasing demands of future applications. An optimized IF amplifier design, using a depletion mode PHEMT technology, is discussed in this article. It will be required to meet the challenging performance requirements of new circuit applications. ■

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Design of a Broadband MIMO RF Transmitter for Next-generation Wireless Communication Systems

The development of a broadband MIMO RF transmitter fulfilling the requirements of the IMT-Advanced wireless communication system is presented. The channel bandwidth of the transmitter is 100 MHz operating in the TDD mode with more than 25 dBm linear output power and 60 dB output power control range. Excellent EVM performance is obtained. The RF transmitter can support up to 6 × 6 MIMO configuration and has been successfully integrated with the MIMO RF receiver and the baseband modules. The average data rate in the field test is approximately 1 Gbps.

The last decade has seen a great burst in wireless communication technologies and their commercial implementations. Until now, the third generation (3G) mobile networks (such as UMTS and cdma2000) and 3.5G mobile networks (such as UMTS/HSPA and CDMA 1xEvDO), which can deliver data rates up to several Mbps to individual users, have been deployed in many countries and adopted by more and more people. However, the pursuit for higher data rates never stops, and a number of wireless technologies are under development to meet future needs: Long Term Evolution (LTE) and WiMAX, both of which can support far higher data rates than existing networks. According to the definition of IMT-Advanced (4G) systems by the International Telecommunication Union (ITU), the next generation network should support data rates up to 100 bps for high mobility and approximately 1 Gbps for low mobility. Unfortunately, neither LTE nor WiMAX can fulfill these requirements. Currently, much work is going on to enhance these exiting standards in order to meet the requirements of the IMT-Advanced system.[1,2] The roadmap of evolutions of various candidates toward the IMT-Advanced system is shown in ***Figure 1.***[3]

To meet the requirement of the data rates defined in the IMT-Advanced system, the channel bandwidth will be up to 100 MHz. Moreover, the MIMO technique is needed to obtain higher channel capacity or spectrum efficiency. When the channel bandwidth approaches 100 MHz, RF designers have to face many challenges, both in the system scheme and the circuit design. It is known that the 6 × 6 MIMO configuration can provide more than 10 bit/Hz in the spectrum efficiency in many typical wireless channels without too much difficulty. In this article, a detailed description of the design of a high performance MIMO RF transmitter with a novel direct-conversion architecture is provided. The RF transmitter operates in the 3.45 GHz frequency band with 100 MHz channel bandwidth and excellent EVM and gain flatness performance. Experimental results are reported. The transmitter can support the 6 × 6 MIMO configuration and has been successfully used in an IMT-Advanced experimental network.

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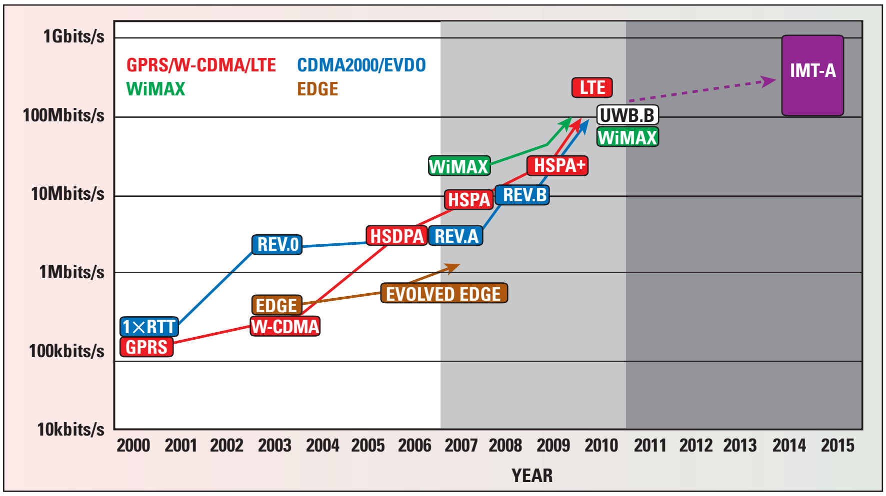
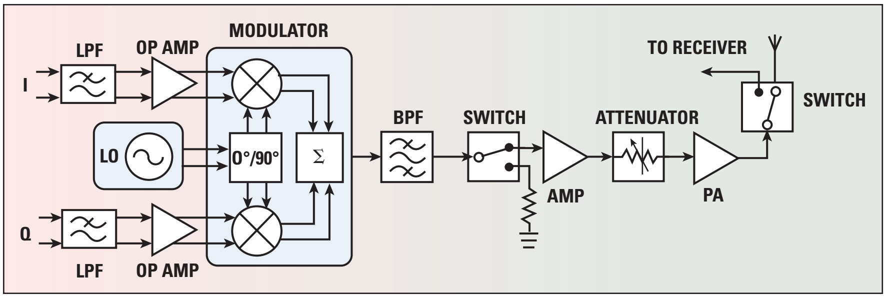
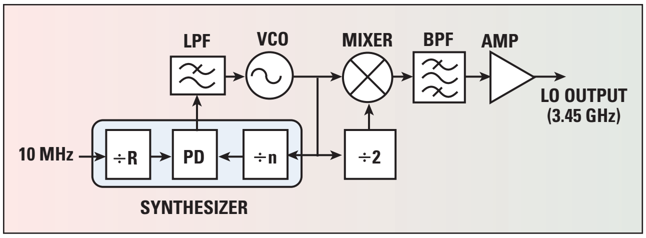


▲ *Fig. 1 Evolution of mobile technologies toward the next generation system.*



▲ *Fig. 2 System diagram of the direct conversion transmitter.*



▲ *Fig. 3 Architecture of the LO generator.*

SYSTEM ARCHITECTURE

Compared to the conventional superheterodyne architecture, the direct-conversion solution has gained more and more attention and applications in various low-cost and compact wireless communication systems.[4-6] Unlike the direct-conversion receiver, which has design challenges still needed to be resolved in order to achieve high performance, it is relatively easier to build a direct-conversion transmitter.[7]

The transmitter reported utilizes a direct-conversion architecture illustrated in ***Figure 2***. This diagram is just an abstract illustration and only essential components are outlined. Other necessary circuits, such as the control circuit, power supply circuit, etc., are omitted for brevity.

The configuration of the PLL is shown in ***Figure 3*** and will be introduced later. As shown, the transmitter utilizes differential analog I/Q inputs, which are used not only to get a better signal-to-noise ratio due to common mode noise rejection, but also to suppress the even order distortions resulting from the nonlinearity of the quadrature modulator. The low pass filters following are used to further reduce the out-of-band emission level and aliasing products.

Next, the operational amplifiers can provide a certain DC offset, required by the inputs of most quadrature modulators. Usually these operational amplifiers should not provide very much gain since the baseband signal level is relatively large and even a few dB gain can saturate the modulator. In this transmitter, the input stage uses operational amplifiers of unit gain. The bandpass filter after the modulator suppresses the spurs resulting from the nonlinearity of the modulator. The transmitter works in the TDD mode, so it is important to switch off the RF amplifier and power amplifier as fast as possible when the receiving period is on, but the bypass capacitors necessary for the stability of the amplifiers always make this unpractical. To prevent the receiver from saturation or jamming by the power amplifier at the start of the receiver period, it is wise to use a RF switch, right before the RF driver amplifier together with the one before the antenna, both of which feature fast switch speed within 60 ns and thus provide enough isolation between the receiving and transmitting period quickly.

The Automatic Power Control (APC) is realized with digital attenuators, which can provide more than 60 dB attenuation control range in 0.5 dB steps. The required output power of the transmitter is 20 dBm. Considering the peak-to-average ratio of the transmitted signal, a MMIC power amplifier capable of 33 dBm output power at its 1 dB compression point is utilized. To further assure the linearity, the output IMD level is tuned and optimized when the output power is approaching 25 dBm.

Usually, special attenuation should be paid in the design of the PLL used in the direct-conversion architecture in order to avoid the LO pulling effect, which can seriously deteriorate the system performance. As briefly shown in Figure 3, the transmitter works on a carrier with an offset from the running frequency of the VCO. The offset is produced using a by 2 frequency divider fed by the VCO; the 3.45 GHz carrier is then obtained by mixing the output of the VCO with this offset. The bandpass filter is used to suppress the unwanted mixing products. The common reference clock of 10 MHz shared by the 6 × 6 MIMO system is produced either by an OCXO whose frequency can be precisely adjusted by the system AFC loop, or by the outside reference source, meanwhile bypassing the inside OCXO. This architecture is very flexible in a prototype system and capable of minimizing the frequency offset between the receiver and the transmitter, which can degrade the system performance especially when using an OFDM modulation scheme. Furthermore, the LO phase noise should be carefully optimized to achieve good system performance.[8]

CONSIDERATIONS OF BROADBAND OPERATION

In the case of broadband operation, the performance of the quadra-

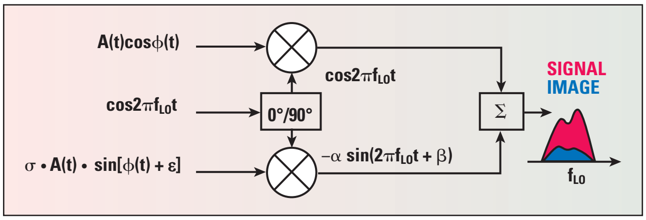


Fig. 4 Simple model representing various unbalanced components resulting in the image product.

ture modulator greatly determines the overall transmitter performance, such as the modulation accuracy indicated by EVM.[9] Compared to the narrowband case, the amplitude and phase unbalance among the differential I/Q input channels becomes more serious and thus brings about the image product, which aggravates the system performance. Furthermore, almost every quadrature modulator in the market requires a DC offset at its differential I/Q input, the unbalance of which makes the elimination of the LO feed through problem more difficult.

IMAGE PRODUCT

The mechanics resulting in the image product can be clarified using a simple model shown in ***Figure 4***. As illustrated, σ and α represent the amplitude unbalance and ε and β represent the phase unbalance. The image lies in the same frequency band as the signal, thus degrading the modulation accuracy. The lengthy development of the output signal equations can be found in the literature.[7] In the case when OFDM is utilized, the up converted signal and its image are shown in ***Figure 5***, in which the individual carrier and its image are drawn in the same color.

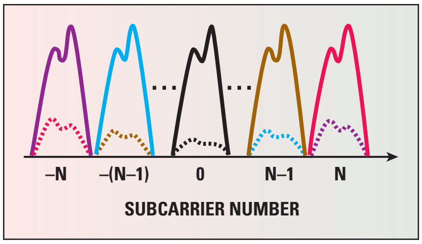


Fig. 5 Up-converted signal and its image when using OFDM.

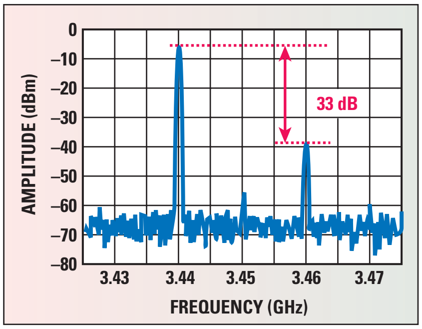


Fig. 6 The unadjusted image suppression.

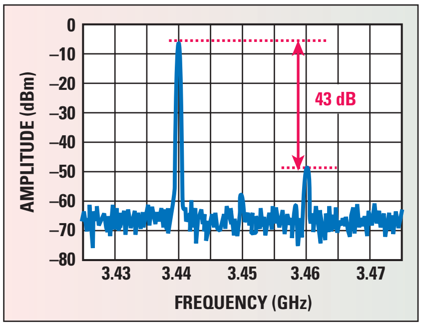


Fig. 7 The adjusted image suppression.

In reality, the amplitude and phase unbalances always exist in the differential I/Q channel, so the image product cannot be completely avoided. Of course, the unbalance of the LO path in the quadrature modulator can also result in the image product, but the influence is relatively smaller. Usually, the image product can be compensated in the baseband processing, such as adaptive adjustment of the amplitude and phase of the I/Q signal as part of the digital predistortion loop, or simply setting a fixed amplitude and phase offset.

In the RF scenario, to alleviate the performance degradation caused by the image product, attention should be paid to the PCB layout process where the differential I/Q channels should be identical in their physical layout. In the tuning process, the level of the image product can be evaluated from the image to signal ratio, which can be measured using a spectrum analyzer. The unadjusted single tone output with its image product is shown in ***Figure 6***. Then, capacitors on the order of several pF can be shunted to ground from the signal traces of the differential I/Q channels. The working of the capacitors results from their non-ideal parameters, which can be modeled as a complex impedance; the magnitude and phase of the baseband signal can then be slightly tuned. After carefully tuning, an image to signal ratio or image suppression below -35 dBc can be achieved throughout the working band. The improved image suppression is shown in ***Figure 7***.

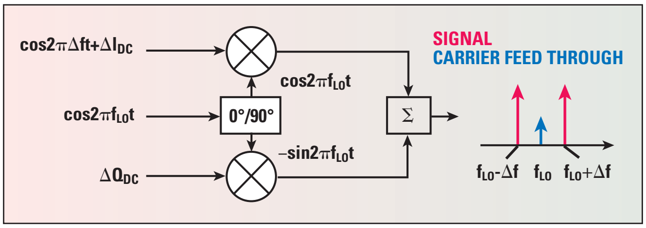


Fig. 8 Simple illustration of the carrier feed-through problem.

CARRIER FEED THROUGH

The carrier feed through results not only from the DC offset unbalance among the I/Q differential inputs, but also from the LO leakage of the quadrature modulator. Usually the LO leakage is not a concern because it is very small in a MMIC quadrature modulator. The mechanics behind the carrier feed-through problem due to DC offset unbalance is illustrated in ***Figure 8***. If a two-tone output is desired, the baseband I/Q signal can then be expressed by the following, with their corresponding DC offset:

I channel signal: $I(t) = \cos 2\pi\Delta f t + \Delta I_{DC}$ (1)

Q channel signal: $Q(t) = \Delta Q_{DC}$ (2)

Output signal:

$$O(t) = (\cos 2\pi\Delta f t + \Delta I_{DC}) \cdot \cos 2\pi f_{L0} t - \Delta Q_{DC} \cdot \sin 2\pi f_{L0} t$$

$$= \frac{1}{2}\left[\cos 2\pi(f_{L0} + \Delta f)t + \cos 2\pi(f_{L0} - \Delta f)t\right]$$

$$\underline{+A\cos(2\pi f_{L0} t + \theta)} \quad (3)$$

$$A = \sqrt{\Delta I_{DC}^2 + \Delta Q_{DC}^2} \quad ; \quad \theta = \tan^{-1}(\Delta Q_{DC} / \Delta I_{DC})$$

According to the above equations, the component underlined in Equation 3 is the carrier feed-through product. The input DC offset levels can be slightly adjusted around the recommended common mode voltage required by the quadrature modulator to greatly reduce the carrier feed-through level. The circuit realizing the tuning function is shown in ***Figure 9***. After iterative adjustment of the DC offset of the I/Q differential input, the carrier feed through to signal ratio or carrier suppression below -50 dB is easily achieved. The tuning result is shown in ***Figure 10***.

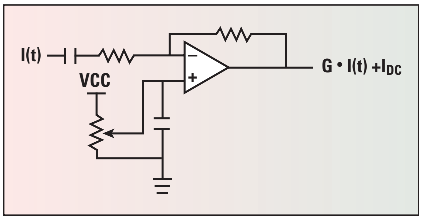


Fig. 9 Circuit realizing the tuning of the DC offset level using an operational amplifier.

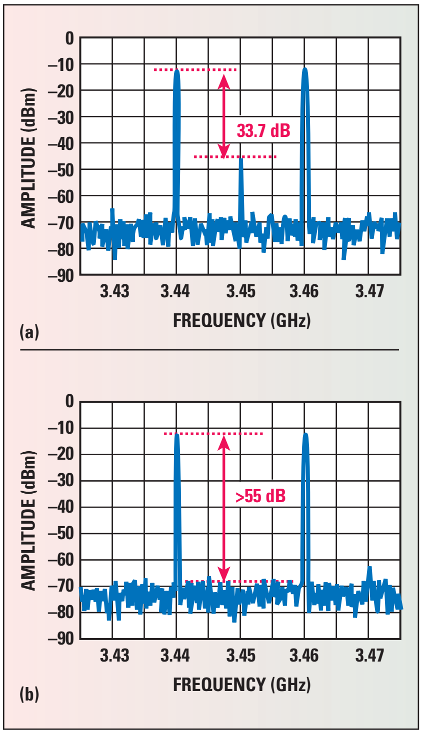


Fig. 10 The unadjusted carrier feed through (a) and the adjusted carrier feed through (b).

MEASUREMENT RESULTS

The transmitter reported in this article is integrated together with the receiver into one single PCB board. The whole 6 × 6 MIMO system consists of six such transceiver modules as well as one control board, which acts as an interface to the baseband serial control signals, and one power supply board providing the DC-DC conversion from 48 to 6 V used by the transceiver. The 6 × 6 MIMO system equipped with six transceiver modules is illustrated in ***Figure 11***.

The transmitter is tested with analog I/Q inputs from a vector signal generator. During the test process, it is convenient to divide the transmitter into three parts: modulator, driver and power amplifier, which can be tuned and tested individually. In the modulator test, the performance that is cared about is the modulation accuracy indicated by EVM. The LO phase noise should first be optimized by adjusting the setting of the synthesizer and the loop filter. After adjustments of the unbalance level mentioned before, the modulation accuracy can then be guaranteed. The resulting LO phase noise figure and the EVM of different modulation type are listed in ***Table 1***, together with the image to carrier ratio and carrier feed through level. Due to the limited capability of the laboratory instruments, the maximum signal bandwidth that can be analyzed is 20 MHz.

A conventional S-parameter test is carried out in the testing of the driver and power amplifier. After careful tuning, the gain fluctuation is kept below 0.5 dB in each stage. An additional IMD test is also carried out to evaluate the nonlinearity of the power amplifier when the output level of 25 dBm is achieved. The results are listed in ***Table 2***.

After individually testing of the different stages, the whole link is connected and tested. The overall performance, indicated by the EVM when the output level achieves the specified 20 dBm, is listed in ***Table 3***.

With help from the Agilent Open Laboratory in Beijing, China, the transmitter was tested with a baseband signal of 80 MHz modulation bandwidth, which was the maximum digital modulation bandwidth available from Agilent at that time. The results are shown in ***Figure 12***. The channel 1 OFDM error summary is shown in ***Table 4***.

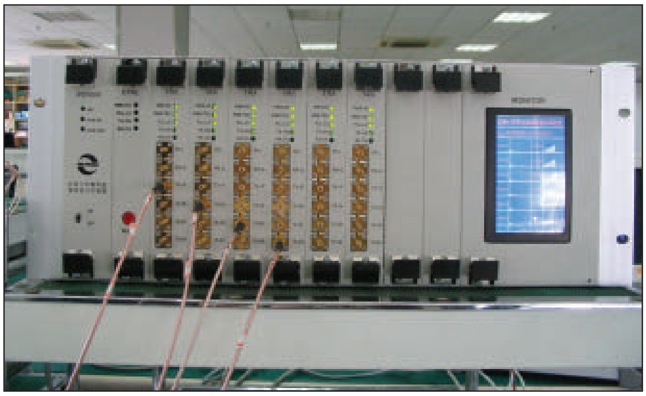


Fig. 11 The 6 × 6 MIMO RF transceiver.

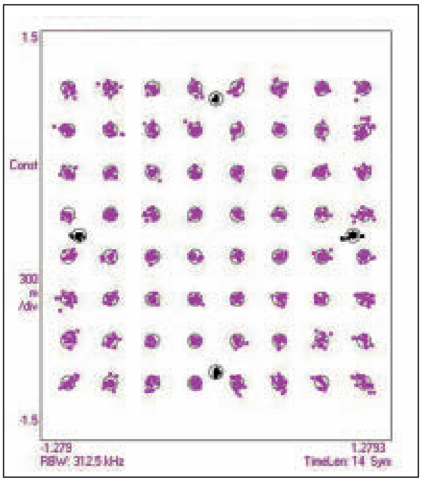


Fig. 12 Constellation diagram.

TABLE I
MEASUREMENT RESULTS OF THE MODULATOR

LO phase noise:	-82 dBc/Hz@1 kHz;-95 dBc/Hz@10 kHz		
	-120 dBc/Hz@10 kHz;-139 dBc/Hz@1 MHz		
EVM & SNR			
Modulation Type (20 MHz)	**EVM (%)**	**SNR (dB)**	**Carrier Suppression**
QPSK	1.5-1.8	35	-50 dB
QAM16	1.1-1.2	35	**Image Suppression**
QAM64	1.0-1.1	35	-42 dB

TABLE II
MEASUREMENT RESULTS OF THE DRIVER AND POWER AMPLIFIER

Driver Part		**Power Amplifier Part**		
Gain (dB)	Return Loss (dB)	Gain (dB)	Return Loss (dB)	IMD3
Max: 6.8	Input: ≤15	Max: 20.2	Input: ≤17	@Pout= 25 dBm
Ripple: 0.2	Output: ≤20	Ripple: 0.5	Output: ≤27	45 dBc

TABLE III

OVERALL SYSTEM PERFORMANCE AT THE SPECIFIED OUTPUT POWER

Modulation Type (20 MHz)	EVM (%)	SNR (dB)
QPSK	1.3-1.6	36
QAM16	1.0-1.3	35
QAM64	1.0-1.1	35

TABLE IV

ERROR SUMMARY

	Ch 1	Ch 2	Avg	Units
EVM	-29.353	—	-29.353	dB
EVM Peak	-19.505	—	-19.505	dB
Pilot EVM	-31.047	—	-31.047	dB
Data EVM	-29.276	—	-29.276	dB
Freq Err	—	—	80.690	Hz
Sym Clk Err	—	—	-0.44206	ppm
CPE	—	—	0.43687	%rms
IQ Offset	-20.361	—	-20.361	dB
IQ Quad Err	-0.74561	—	-0.74561	deg
IQ Gain Imb	-0.06951	—	-0.06951	dB
Cross Pwr	—	—	—	
Sync Corr	0.92890	—	0.92890	

CONCLUSION

The implementation of a broadband direct-conversion transmitter is reported. The considerations of system architecture are discussed in detail and the issues regarding the broadband working are illustrated. The transmitter was integrated in a 6 × 6 MIMO wireless prototype system capable of data rates up to 1 Gbps in a low mobility scenario. ■

ACKNOWLEDGMENT

This work was supported in part by NSFC under Grant 60621002 and 60702027, 60921063 and in part by the National 973 project 2010CB327400, by the National High-Tech Project under Grant 2008AA01Z223, 2008ZX03005-001 and 2009AA011503.

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[Continued from pg. 10]

still further behind at around 250 ms. Latency is a useful measure of the all-important round-trip speed that is vital to the user appeal of real-time, interactive applications. Think of multi-player gaming and media-rich social networking for an idea of just some of the applications that will be enhanced by LTE-Advanced. ***Table 1*** shows the performance characteristics for each technology.

Also taking centre stage is the ability of LTE-Advanced to leverage advanced topology networks. Reflecting the ITU's vision for IMT-Advanced, Release 10 caters for seamless interworking with a jigsaw of radio access systems. Macro-, micro-, pico- and femto-cells all figure in a heterogeneous network environment, covering cell sizes from tens of kilometres to just a few meters.

As long ago as 2003, ITU-R articulated its long-term strategic vision for IMT-Advanced: a global platform to build the next generation of interactive mobile services, encompassing fast data access, unified messaging and broadband multimedia.

The ITU's formal process of identifying technology candidates for this, a new wireless generation moved forward with an invitation (issued in March 2008) for the submission of proposals for candidate radio interface technologies for the terrestrial components of IMT-Advanced.

In October 2009, the 3GPP Partners made their formal submission to the ITU, proposing LTE Release 10 and beyond—'LTE-Advanced'—as a candidate for IMT-Advanced. Of the five other technology candidates that were submitted in parallel, some were technically identical, leaving just two main candidates. Self-evaluation results in 3GPP have shown that LTE-Advanced meets—or in some cases exceeds—all requirements of ITU-R.

After the expected approval of LTE-Advanced specifications in December 2010, it is anticipated that work on Release 10 will be effectively completed by mid-2011. Following final approval of LTE-Advanced by the end of 2011, this will give vendors and operators a clear target to start building 4G networks. While estimates vary, this timeframe points to initial LTE-Advanced deployments around 2015.

As proposed by the 3GPP Partners, LTE-Advanced is the next technological iteration in a continuum of wireless standardisation at a global level that spans almost three decades.

As with GSM and W-CDMA/HSPA/LTE currently, there will be a long period of co-existence between 3G and 4G systems—maybe for two decades or more. As 2010 draws to a close, there are a growing handful of commercially deployed LTE networks based on 3GPP Release 8, with a flood of further launches expected during 2011 and 2012. These '3.9G' networks are already giving customers an early taste of the possibilities of ultra-high speed mobile broadband, and—tantalisingly—a glimpse of our true 4G future. ■

Jean-Pierre Bienaimé *was elected Chairman of the UMTS Forum in 2003. Since joining France Telecom (FT) in 1979, he has served as Advisor to the General Director of Moroccan Telecommunications in Rabat, Director of Marketing and Product Development for international business networks and services at FT, Director of Business Development and Subsidiaries at France Cables and Radios, Chief Executive Officer of Nexus International and VP International Development at France Telecom Mobile. After the purchase of Orange by FT, he was appointed VP Group Technical Support. He is currently Senior VP, Strategy and Communications at Orange Wholesale.*

BAW Innovation Helps WiFi and 4G to Happily Coexist

Channels allocated for wireless services are increasingly being crammed together on the frequency band while the price per MHz of spectrum remains quite high at auction. This is especially at issue for 4G signals, which are commonly located at frequencies adjacent to existing WiFi and Bluetooth channels and therefore suffer from mutual interference issues. As such, operators and device makers need interference mitigation solutions. One such solution arises out of an advanced filter technology known as bulk acoustic wave (BAW) that has emerged over the past few years. This article will investigate the driving forces behind the need for better filtering for 4G applications and discuss an effective hardware-based solution.

As demand for broadband services grows, the trend towards wireless is well-established. In emerging markets, such as China and India, build-out of wireline infrastructure will be too expensive, time consuming and logistically complicated. In established economies, mobility is the "killer app" that drives the growth in demand for wireless devices and applications. But wireless communications require spectrum, which is—in economic terms—a scarce resource; this has three key implications.

First, a limited supply coupled with strong demand will naturally lead to high prices. This has been evident in the billions of dollars generated at auction for the rights to these bands. For example, this year's 4G auctions in India generated \$8.2 B. The popularly quoted revenue figure for the India BWA auction is \$5.5 B (USD), which doesn't take into account the additional \$2.7 B paid by state-run operators BSNL and MTNL. Without having to actually participate in the auction, these two firms were guaranteed spectrum at a price equivalent to the winning bid. This equated to an average of \$6.2 M per MHz of spectrum, with that price per MHz per capita peaking over \$1.15 in Mumbai and Delhi.

Josh Raha
TriQuint Semiconductor, Hillsboro, OR

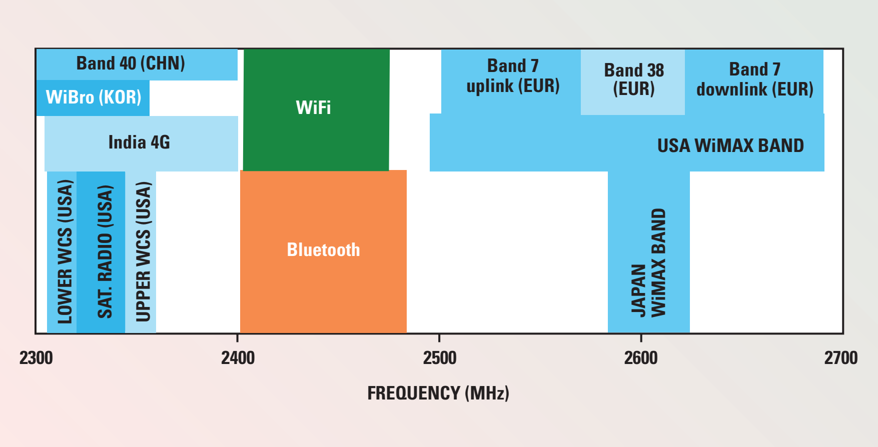


Fig. 1 Allocated frequency bands.

Second, high demand for a scarce resource motivates a supplier to produce as much of that resource as possible. While it is not possible to "produce" more electromagnetic spectrum, it is possible to "refarm" existing spectrum. To that end, regulatory bodies worldwide are working to find more channels to put into use, with the assumption that interference or other considerations will be handled by the marketplace. The FCC, with the support of President Obama, has released The National Broadband Plan this year. Amongst other recommendations, the document calls for 500 MHz of new spectrum to be made available for broadband within 10 years, 300 MHz of which should be allocated for mobile use within five years. To the Broadband Plan's credit, it does recognize that interference issues are lurking, but it does not offer solutions. Meanwhile, many bands allocated for 4G data services happen to be immediately adjacent to the International ISM band, the unlicensed band that runs roughly from 2.4 to 2.5 GHz and is used worldwide for Wireless LAN and Bluetooth signals. Another example in the US is the satellite radio band, which sits right in the middle of WCS spectrum with no guard-band defined by the FCC (see ***Figure 1***). As band assignments get tighter, interference issues multiply.

Third, scarcity leads owners of a resource to use it as efficiently as possible. There are two main approaches to achieving spectral efficiency, both equally important. The first is the use of better modulation techniques, allowing an operator to pack more data into a given channel. The second is to find ways to stretch the use of purchased spectrum to its edges—to use an entire 20 MHz channel for data rather than give up 5 MHz on each side as guard-bands.

Let us take a moment to consider guard-bands. A guard-band is the slice of frequency between two channels, a sort of "no-man's-band" where both parties gradually roll off their transmission power. In some cases, guard-bands are built into the operator's license; the FCC might designate 10 MHz between channels that cannot be used. More commonly, the regulatory body tends to leave the definition of the guard-band to the operator. The requirements for out-of-channel transmissions will be set by, say, the FCC, and the license holder is free to use as much spectrum as he wants, providing he meets those rules. Guard-bands, naturally, represent wasted spectrum, and operators are intent on minimizing them.

Arising out of these implications is a clear picture: 4G operators are spending a great deal of money on channels that happen to lie very close on the frequency spectrum to interfering signals, but have no built-in guard-bands. The standards upon which 4G data services will be delivered are WiMAX, LTE and TD-LTE. These are extremely similar in terms of transmission pattern (all use the OFDMA modulation scheme); and, more importantly, there is a significant overlap in the frequencies that each will use. As

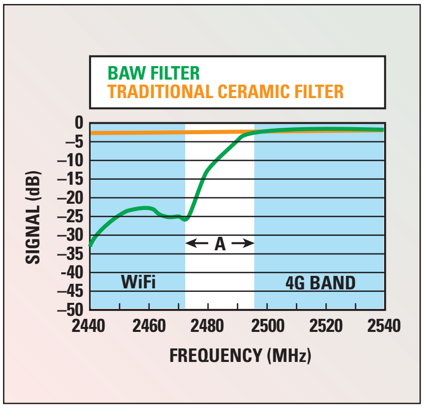


Fig. 2 Example filter performance.

such, for the purposes of this discussion, they can be referenced collectively as "4G". While there are sub-2 GHz bands set aside for LTE (most notably, the 700 MHz "digital dividend" spectrum), a significant proportion of 4G services will be delivered in what the author calls the 4G bands: 2.3 to 2.4 and 2.5 to 2.7 GHz (see Figure 1).

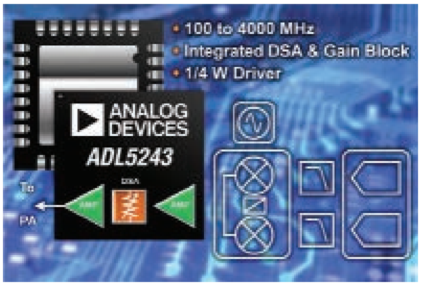
Operators find this spectrum attractive for three main reasons. First, it has relatively good RF propagation characteristics. Second, it is largely unused, meaning there are bigger chunks of consecutive spectrum available here than elsewhere in the band. Third, we are seeing some level of global harmonization, with governmental regulators in India, China, Japan, America and Europe all offering this spectrum for 4G services. Alas, nothing is perfect and there is a key drawback, discussed earlier: nestled amongst these attractive 4G bands are nasty interferers in the form of WiFi, Bluetooth and satellite radio signals.

This issue is perhaps most apparent and acute in the increasingly popular "personal hotspot" devices. These are wireless LAN access points that take a 4G signal from the service provider and convert it to WiFi in order to share the 4G connection across multiple WiFi-enabled devices. Because they transmit and receive 4G and WiFi signals at the same time, these personal hotspots have the greatest risk for mutual interference issues.

In order to address this issue while minimizing wasted spectrum in the guard-bands, operators and device makers have turned to advanced filtering technologies. A filter with a steep skirt—one that rolls off quickly from pass-band to rejection band—becomes increasingly important in cases like these (see ***Figure 2***). Surface acoustic wave (SAW) filters have traditionally served this purpose. However, as frequencies rise to the 2.3 to 2.7 GHz range, SAW performance starts to decline. BAW filters are the natural evolution of acoustic wave technology for higher frequency bands, and become extremely attractive as frequencies grow to 2.3 GHz and above.

For example, TriQuint Semiconductor's BAW technology has been applied to a trio of filters specifically designed to mitigate the interference issues between the ISM and 4G bands. One filter in this family passes the ISM band while effectively rejecting signals above 2.5 GHz. One application for this part would be as a WiFi pass-band filter. Notch filters are used to knock down WiFi/BT signals and pass the 4G signals with low insertion loss both above and below the ISM band. In addition to excellent electrical performance, BAW filters can be extremely small with very low profile packaging. Filters are available in 1.3 × 1.7 mm packages with a profile of less than 0.5 mm. Clearly, this supports the long-term trends toward ever-smaller, full-featured mobile devices. BAW also exhibits excellent power handling capabilities, with WiFi notch filters typically withstanding +28 dBm (continuous wave). Used in various combinations depending on the end application, these BAW filters are good examples of how companies are simplifying RF connectivity with new hardware technology that resolves real-world problems like the interference and band adjacency issues 4G operators now face.

Demand for 4G will continue its path of explosive growth and it is conceivable that, moving forward, every scrap of spare spectrum will be prized as a vehicle for delivery of lucrative broadband data services. As useable spectrum bands continue to press together, innovative filtering solutions like BAW technology will continue to help operators make better use of their spectrum and realize more complicated and compelling consumer devices. ■



IF and RF DVGAs for Next Generation Wireless Systems

Analog Devices has introduced four new Digital Variable Gain Amplifiers (DVGA): ADL5201 and ADL5202, which are IF DVGAs; and ADL5240 and ADL5243, which are RF DVGAs. These new DVGAs will enable the design of smaller base stations with multiple carrier capabilities for next generation wireless systems such as Long Term Evolution (LTE) systems.

The ADL5201 and ADL5202 DVGAs are optimized for use in IF sampling receivers. The ADL5201 (see ***Figure 1***) is suitable for single-channel receivers while the dual-channel ADL5202 is suitable for use in main and diversity or MIMO receivers. For the most common IFs used between 70 and 300 MHz, both these devices exhibit minimum amplitude variations over frequency, allowing designers to choose an optimum IF for their design. With a gain range of 31.5 dB (-11.5 to 20 dB), these devices can be used to expand the dynamic range of high performance IF sampling receivers. In addition, a gain-control step size of 0.5 dB ensures that the full input range of the analog-to-digital converter (ADC) can be fully utilized.

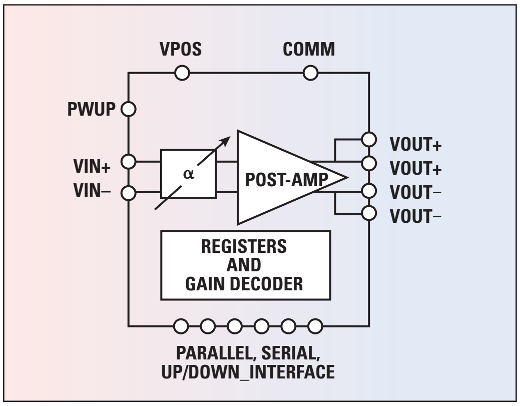


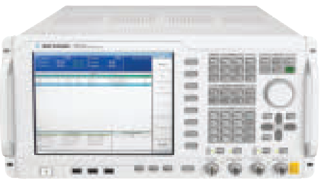
Fig. 1 ADL5201 block diagram.

The ADL5201 and ADL5202 DVGAs feature novel and flexible gain control interfaces. In addition to parallel and serial interface options, a novel up/down mode is also available. When operated in parallel mode, the 6-bit gain code can either be latched into a register or the register can be made transparent resulting in the gain following the code on the six gain control pins. In serial mode, the gain is set by clocking serial data into the devices' serial to parallel interface (SPI), which also has a read back mode. In the serial mode, there is an additional fast attack mode where rather than setting the gain to a specific level, the gain can be changed by 2, 4, 8 or 16 dB step increments or decrements. The gain can also be changed using an up/down interface. This allows the gain to be incremented up or down in steps of 0.5, 1, 2, or 4 dB. The gain control interfaces of the

[Continued on pg. 34]

Analog Devices Inc.
Norwood, MA

Mobile Communications: Company Showcase



LTE Base Station Emulator

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Agilent Technologies has a new LTE Base Station Emulator designed to speed development and verification of LTE user equipment. The PXT Wireless Communications Test Set (E6621A) is a powerful, common hardware test platform that is used across the LTE development lifecycle. Agilent's PXT test set represents a significant breakthrough in testing handsets and data modems for all planned LTE deployments. The new PXT is the basis for a Signaling Conformance Test suite that will run TTCN-3 scripted test cases validated by PTCRB, the North American regulatory agency responsible for cellular device compliance. For more information, visit www.agilent.com/find/PXT.

**Agilent Technologies Inc.,
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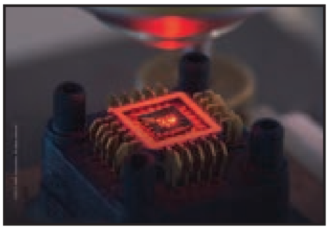

AWR Visual System Simulator 2010

VENDORVIEW

AWR's Visual System Simulator™ software (VSS) is ideal for communication system design. The 2010 release introduces new capabilities that increase productivity for RF engineers, including time delay neural network (TDNN™) advanced amplifier behavioral models for capturing memory effects and a new phased-array element for radar designs. Enhancements have also been made to the VSS's RFA™ architect tool. New communication models and signal processing blocks have been added as well, including turbo decoders for 3G/4G standards. AWR Connected™ for Rohde & Schwarz's WinIQSIM2 waveform generation software integrates with VSS to ensure test & measurement source signals are identical to those within VSS. This additional module to VSS supports all communications standards and their variants. Visit www.awrcorp.com and AWR.TV to learn more about VSS as well as to download a free 30-day trial.

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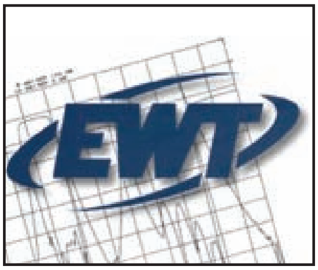

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Eastern Wireless TeleComm Inc. (EWT) specializes in custom design and manufacture of RF and microwave filters and filter-based products for mobile communications applications. The staff encompasses over 50 years of combined experience in the design, development, and high volume manufacturing of cavity and waveguide filters to 50 GHz and lumped element filters up to 10 GHz. All of the company's products are manufactured to strict guidelines set forth in its ISO-9001 compliant quality system, ensuring the highest level of product performance and reliability. Through utilization of the company's in-house machining capabilities and state-of-the-art manufacturing processes, rapid turn-around is standard. EWT can develop any cavity, lumped element, waveguide, or wireless filter to meet your needs. Visit us online to request your quote today.

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Compact Communications Amplifier Module

The PCM3R3SCO (SKU 7091) is one of a series of communications amplifiers. This 900 MHz compact linear power amplifier is one of a series that covers 450, 700, 800, 900, 1800, 1900 and 2100 MHz bands and is suitable for use in commercial LTE networks and public safety wireless communication applications. The amplifier employs latest generation LDMOS device technology and is highly efficient. The amplifier has an advanced built-in predistortion engine with wide instantaneous correction bandwidth, which ensures low distortion and wide dynamic range operation. Efficiency of up to 40 percent is possible in this series where the high performance is achieved by employing advanced RF design techniques. Empower RF's ISO9001 Quality Assurance Program assures consistent performance and the highest reliability.

**Empower RF Systems Inc.,
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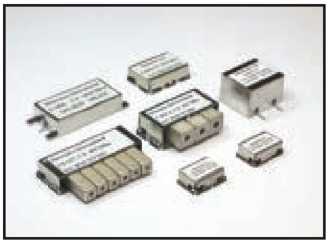

Mobile Communications

NIC's mobile communication products include filters, diplexers and integrated assemblies used in the transmit-receive chain as well as for harmonic suppression. Different technologies such as LC, ceramic and cavity are used based on the specifications. Custom designs are available to support specific requirements. Features include: cavity filter designs with low insertion loss and high power handling capability; custom high Q ceramic filter designs offering high performance in a small package configuration; integrated solutions providing improved performance while significantly reducing the size and simplifying customer procurement; and standardized products to offer low cost solutions. NIC is a global manufacturer of custom RF and microwave components.

**Networks International Corp.,
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RF CMOS Solutions

Peregrine's new PE42662 SP6T features six fully symmetric transmit ports allowing any combination of transmit/receive for 3G front-end modules in multi-band GSM/EDGE/W-CDMA handsets. The UltraCMOS™ switch delivers high linearity and exceptional harmonic performance: 2fo/3fo = -40 dBm at 35 dBm TX (900 MHz) and 33 dBm TX (1900 MHz); IIP3 of +68 dBm at 50 Ohm; low transmit insertion loss of 0.5/0.6 dB (900/1900MHz); and 38 dB/33 dB isolation (900/1900 MHz). The switch handles maximum +35 dBm input power with world-class ESD tolerance of 4000 V HBM at ANT port, and 2000 V HBM on all ports.

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Mobile Communications: Company Showcase

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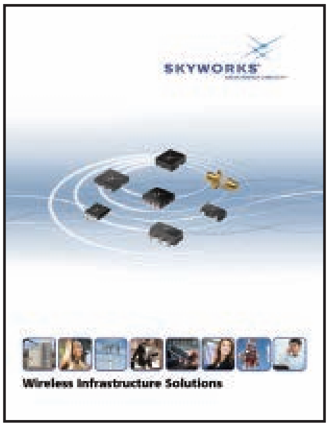

Wireless Infrastructure Solutions

VENDORVIEW

Skyworks' broad portfolio of RF/microwave products includes solutions for today's demanding wireless infrastructure systems including cellular base stations, WiMAX access points, land-mobile radio systems and point-to-point radio links. Skyworks offers components and subsystems from the antenna connection to the baseband output, including low noise amplifiers, power amplifiers, general-purpose amplifiers, mixers, modulators, demodulators, phase shifters, switches, attenuators, detectors, directional couplers, hybrid couplers, power splitters/combiners, ceramic filters and resonators, plus discrete control components including PIN, tuning varactor, Schottky diodes and chip attenuators.

Skyworks Solutions Inc.,
Woburn, MA (781) 376-3000, www.skyworksinc.com.

RF Probe

INGUN has announced an innovative breakthrough in modern RF probing technology with the introduction of the HFS-835 series RF Probe with integrated attenuator. It is said to be the first spring-loaded RF probe with an integrated attenuation module. The 50 Ω RF probe has a broadband frequency range of DC to 3 GHz. Front plungers and inner conductor tip-styles can be tailor-made to accommodate contacting of several different PCB land-pattern shapes as well as selected RF connector types. Applications include virtually any kind of production-line test with RF test applications, e.g. the general telecommunication market, WiFi boards, Bluetooth applications and the automotive market. It is especially suitable for test fixtures with space restrictions or to prevent shear forces to the probe due to externally connected attenuators.

INGUN Prüfmittelbau GmbH,
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www.ingun.com.

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IF DVGAs are designed to allow simple interfacing to ADCs from Analog Devices, such as the AD9643, which features over range detection. When the ADC is connected to an ADL5201 or ADL5202 up/down interface, over range detection at the ADC's output port starts decreasing the gain until the ADC is no longer in an over drive condition.

The ADL5201 and ADL5202 provide outstanding linearity, with an output third-order intercept point (OIP3) of better than +50 dBm at the high end of the gain range at IFs up to 150 MHz. Another key feature of these devices is the low power mode, which reduces the supply current by 25 percent compared to the standard operating mode. Operating in the low power mode results in a moderate drop in linearity, but the mode can be switched on and off based on dynamic conditions in the receiver. For example, under normal receive conditions the low power mode could be used with the higher linearity mode turned on only when large in band blockers are present.

The ADL5240 and ADL5243 are high-performance RF DVGAs that operate over a broad frequency range of 100 MHz to 4 GHz. The ADL5240 integrates a DSA with a broadband, fixed-gain amplifier. The amplifier is internally matched and has a broadband gain of approximately 19.5 dB. The 6-bit DSA has 31.5 dB gain-control range, 0.5 dB step size and ±0.25 dB step accuracy over the entire frequency range. The DSA attenuation can be controlled by using either a parallel or serial interface mode. The DSA and amplifier in the ADL5240 can be wired for the attenuator to drive the amplifier, for transmit applications, or for the amplifier to drive the attenuator, for receive applications. The ADL5240's +38 dBm OIP3 and 3 dB noise figure make the device attractive for both receiver and transmitter signal paths.

The ADL5243 provides an even higher level of integration. Along with a broadband amplifier and a 31.5 dB DSA, the ADL5243 includes a second amplifier. This allows the device to be configured in an amplifier-DSA-amplifier component lineup. When the three components are connected in the aforementioned configuration, they provide a cascaded gain of 29 dB when the DSA is set to minimum attenuation. The ADL5243's final stage amplifier is designed to deliver highly linear output power with OIP3 of +41 dBm and is capable of driving directly into a base station power amplifier. Like the ADL5240, the DSA attenuation in the ADL5243 can be controlled either by a parallel or serial interface mode and support 0.5 dB step size with ±0.25 dB step accuracy.

The newly introduced RF and IF DVGAs from Analog Devices provide significant integration advantages with reduced system and cost complexity and will support and enable small footprint designs for next generation wireless systems.

Analog Devices Inc., Wilmington, MA (978) 658-8930, www.analog.com.

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RFMD®

RF3482
WiFi, Low Band Front End Module

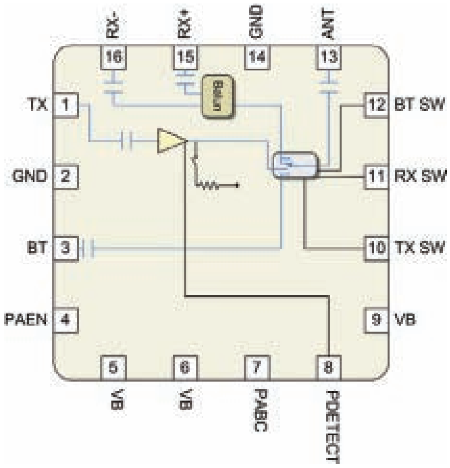
RFMD® now offers the RF3482, a single-chip, integrated front end module (FEM) developed for high-performance WiFi applications in the 2.4 to 2.5 GHz ISM band. This FEM addresses the need for aggressive size reduction for a typical 802.11b/g/n front end design and minimizes the number of components required beyond the core chipset. The RF3482 features an integrated b/g/n power amplifier, power detector, receive (Rx) balun, and some transmit (Tx) filtering. It is also capable of switching between WiFi Rx, WiFi Tx and *Bluetooth*® Rx/Tx operations. The device comes in a 3 x 3 x 0.45 mm, 16-pin package and meets or exceeds the RF front end needs of 802.11b/g/n WiFi RF systems.

SPECIFICATIONS

Part Number	***Functionality***	***IEEE 802.11 Type***	***11g/n P_{OUT} (dBm)***	***11b P_{OUT} (dBm)***	***11b/g/n Gain (dB)***	***11g/n EVM (%)***	***V_{CC} (V)***	***11g/n Operating Current (mA)***	***11b Operating Current (mA)***	***Package (mm)***
RF3482 *	PA, SP3T, Rx Balun, 2170 MHz and 2 Fo Filter	b/g/n	16.0	20.5	33.0	3.0	3.3	135	170	QFN 3.0 x 3.0

*** Power Detector Coupler**

RF3482 Block Diagram



FEATURES

- Single module radio front end
- Single voltage supply 2.7 to 5.5 V
- Integrated 2.5 GHz b/g/n amplifier, Rx balun, and Tx/Rx switch
- P_{OUT}=16 dBm, 11g, OFDM at 3.0% EVM and P_{OUT}=20.5 dBm, meeting 11b mask
- Applications include IEEE802.11b/g/n WiFi; single-chip RF front end module; 2.5 GHz ISM bands; WiFi systems; portable battery-powered equipment; optional *Bluetooth*® sharing of single antenna port

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