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April 2008

Microwave Journal

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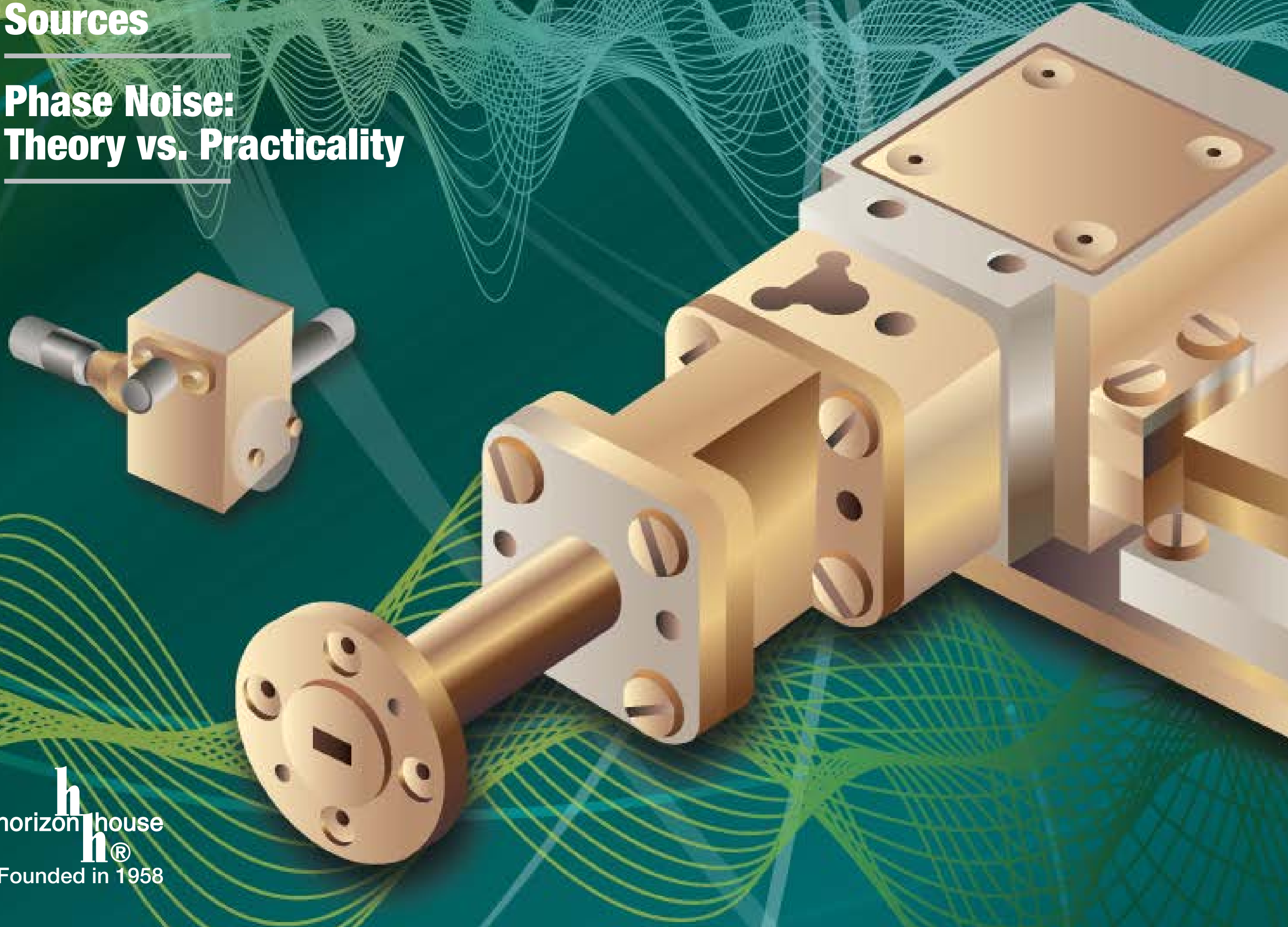
Amplifiers and Oscillators

Similarity Considerations for Varactor Multipliers

Fifty Years of Frequency Sources

Phase Noise: Theory vs. Practicality



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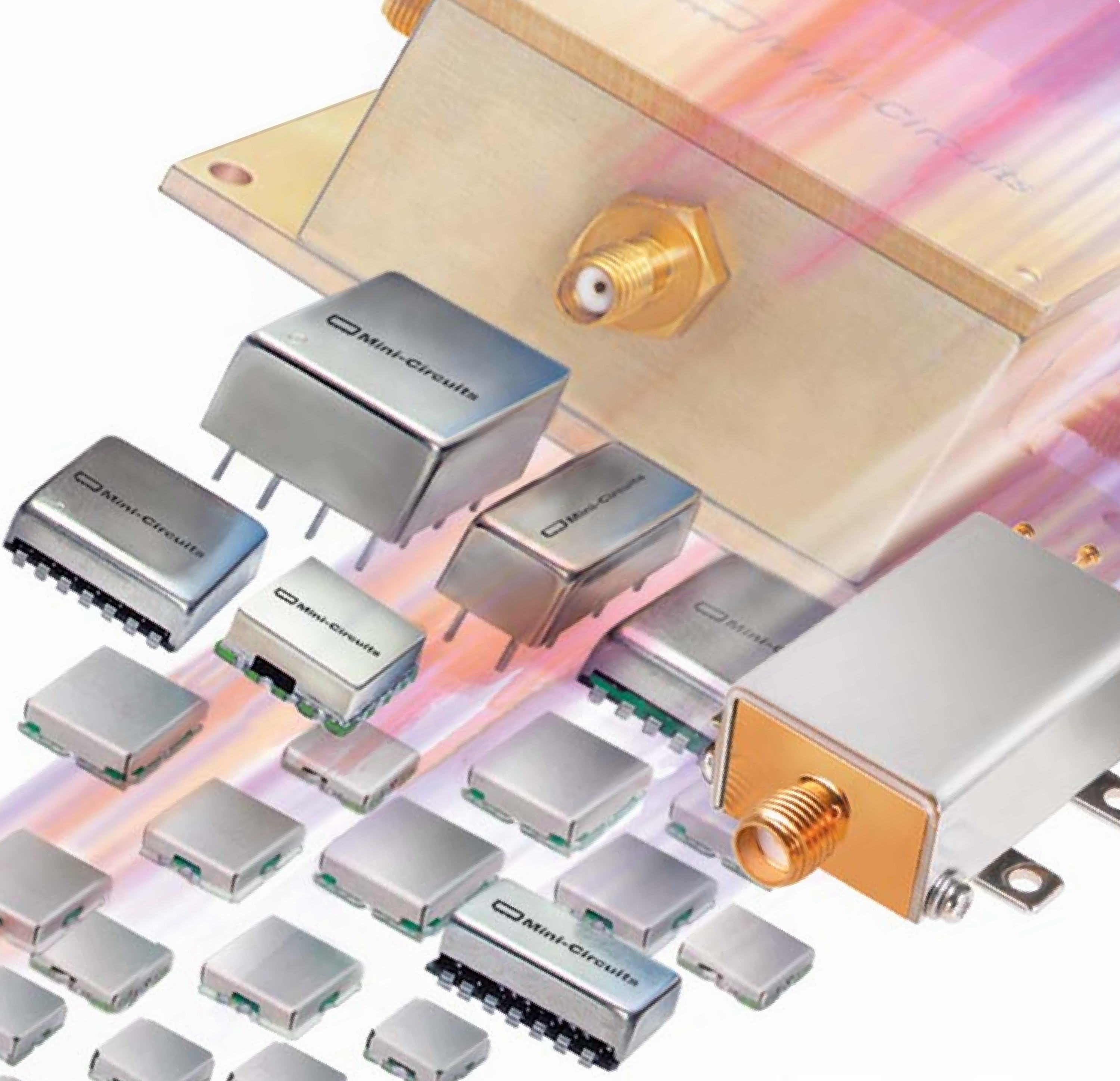
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FEATURES

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STAFF

Executive Editorial Office:
685 Canton Street, Norwood, MA 02062
Tel: (781) 769-9750
FAX: (781) 769-5037
e-mail: mwj@mwjournal.com

European Editorial Office:
46 Gillingham Street, London SW1V 1HH, England
Tel: Editorial: +44 207 596 8730 Sales: +44 207 596 8740
FAX: +44 207 596 8749

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A Dual-band-reject Design for a Cross Semi-elliptic Monopole Antenna
Wen-Shan Chen, Chi-Huang Lin and Mao-Kai Hsu, Southern Taiwan University of Technology

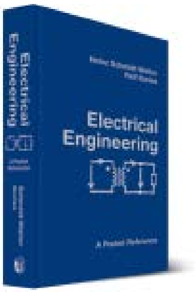
A Novel 3D CP-FDTD Thin Slot Algorithm for Shielding Analysis
Xu Li, Ji-hui Yu, Quan-di Wang and Yong-ming Li, Chongqing University

Microwave Point-to-point Systems in 4G Wireless Networks and Beyond
Harvey Lehpamer, HL Telecom Consulting

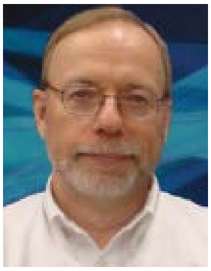
Phase Diversity and Optimal I/Q Signal Combining Methods on a UHF RFID Reader's Receiver
Byung-Jun Jang, Kookmin University

Expert Advice

Microwave Journal asks a noted industry expert to provide commentary related to the month's editorial theme. Respond with comments and win a complimentary copy of *Electrical Engineering: A Pocket Reference* from Artech House (see www.mwjournal.com for details). Come join the discussion, happening now.



April: Dan Nehring, VP of engineering, Frequency Control Products, Valpey Fisher Corp., talks about considerations when specifying frequency control modules and oscillators to optimize system jitter and phase noise.



Executive Interview

In this month's executive interview, **Olle Westblom**, managing director of Sivers IMA AB, provides insight into the workings of an independent Swedish company that has been making advanced microwave oscillator products for over 50 years. He focuses on the development of ultra-wideband VCOs, the use of MMIC technology and the virtues of collaboration.

Event

MWJ editor, Patrick Hindle, reports on RF/microwave exhibitor news from CTIA Wireless 2008, Las Vegas, NV Convention Center, April 1–3.

CTIA WIRELESS 2008

The Old Order Changeth...

(Reprinted from the March 1994 issue of *Microwave Journal*)

Harlan Howe, Jr.

I am very pleased to announce that Mr. Frank Bashore has joined the *Microwave Journal* as Technical Editor. Frank replaces Martin Stiglitz, who has stepped down from that position after nine years of faithful service. Martin started a second career with the *Microwave Journal* in 1985, after retiring from the laboratory at the Hanscom Air Force Base in Massachusetts. Since that time he has been the principal technical contact with the many authors who have published with the *Microwave Journal*. Martin solicited papers, served in our many review meetings, wrote special reports, was the principal author of our book reviews and did almost all of our technical editing. His wit, insight and famous, or perhaps infamous, red pen will be missed.

Frank Bashore is also starting a second career, as have I, after years of experience as a design engineer and engineering manager at a number of prestigious companies in our industry. After an apprenticeship with a number of early microwave companies in New Jersey, Frank moved to New England to assume a series of managerial positions at Frequency Sources Inc. which subsequently became part of Loral Corp. From Frequency Sources, he went to Anzac Division of Adams Russell as director of Custom Products. He left Anzac briefly to work for Microsonics Inc. as Director of Operations, only to return to Anzac as Engineering Manager, which is when I first had the opportunity of working with him. On the basis of that experience, I am confident that Frank will make a positive and continuing contribution to the *Microwave Journal*. After the acquisition of Adams Russell by M/A-COM, Frank became the Manufacturing Manager in the M/A-COM Microelectronics Division prior to his joining *Microwave Journal*.

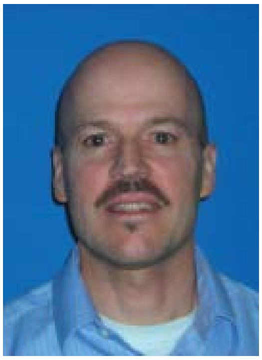
I look forward to working with Frank in the months and years ahead. Welcome aboard!

... Once Again

CARL SHEFFRES

After 14 years, thousands of articles edited and countless product demos, Frank Bashore is retiring as Technical Editor of *Microwave Journal*. Frank became a staple at the MTT-S IMS and other events over the years, meeting with company representatives to discuss their latest products and surveying the landscape for the latest technical innovations. He served as the main contact for our many contributing authors, working with them to insure that their articles were published with the high standards expected in *MWJ*.

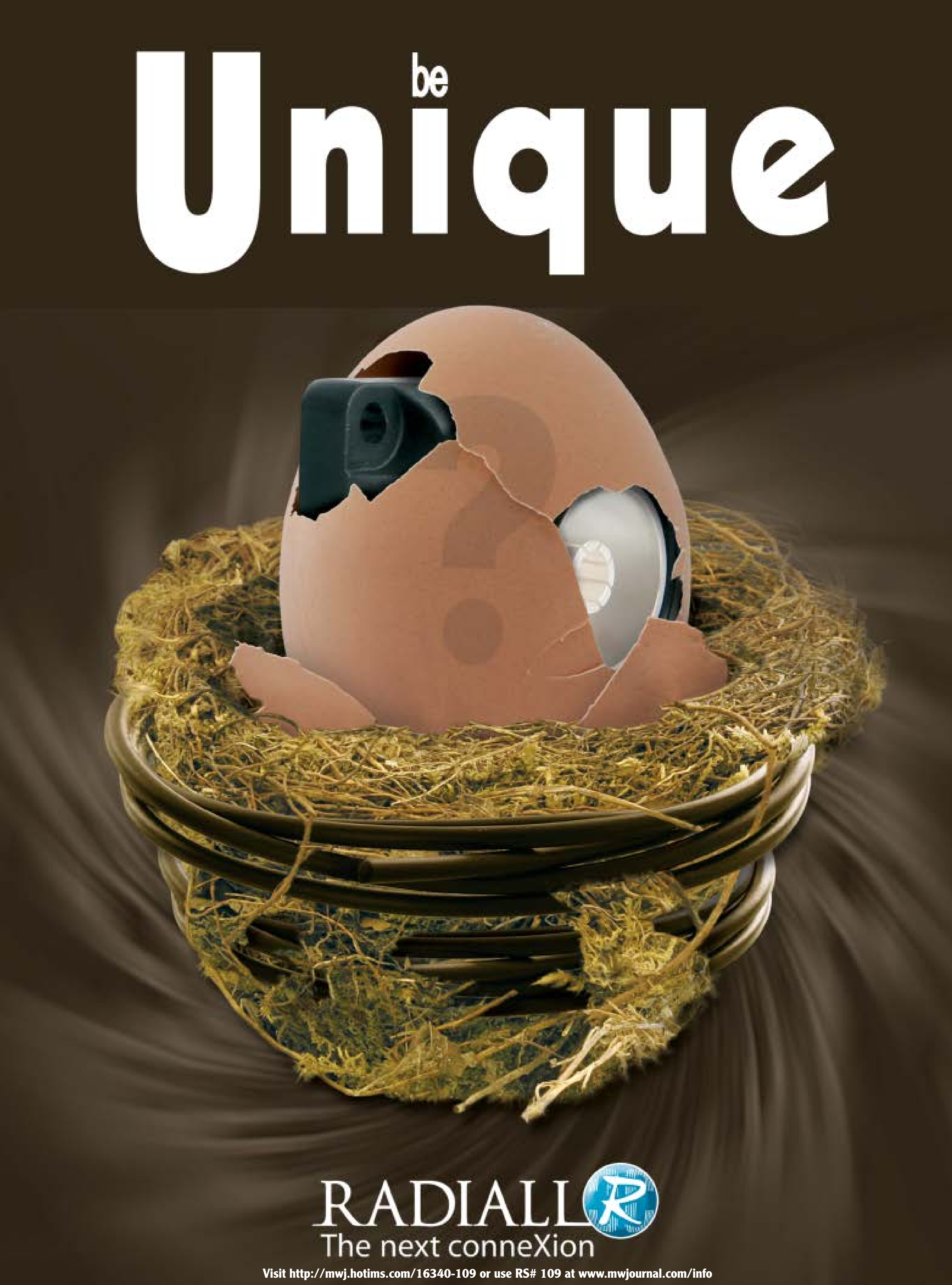
I've had the good fortune to have worked with Frank for all of these years, gaining a genuine respect for his talent, work ethic and good nature. We've travelled together to some exotic and some not so exotic destinations, played many rounds of golf and shared many lasting memories. His daily presence will be missed by all of us at *MWJ*, but we're comforted in knowing that he will remain on our editorial review board, along with Harlan Howe, so they will both still grace us with their monthly presence. We wish Frank all the best in his well deserved retirement.



Frank's worthy successor is Patrick Hindle. I'm pleased to announce that Pat has joined *Microwave Journal* as Technical Editor. Pat brings more than 20 years of experience in the RF/microwave industry to his new position. He has held managerial positions at Raytheon, Alpha Industries/Skyworks, MIT and most recently at M/A-COM Tyco Electronics. His technical background and marketing expertise are assets that will serve our authors, readers and advertisers well in today's dynamic publishing environment. Pat will contribute to the print magazine and our numerous electronic media products, bringing the latest product and technology innovations to our valued readers.

I've had the pleasure of working with Pat on various *MWJ* projects during his years at M/A-COM and have come to appreciate his professionalism, creativity and pleasant demeanor. I look forward to working with him in the years ahead.

Pat can be reached at phindle@mwjournal.com. Welcome aboard!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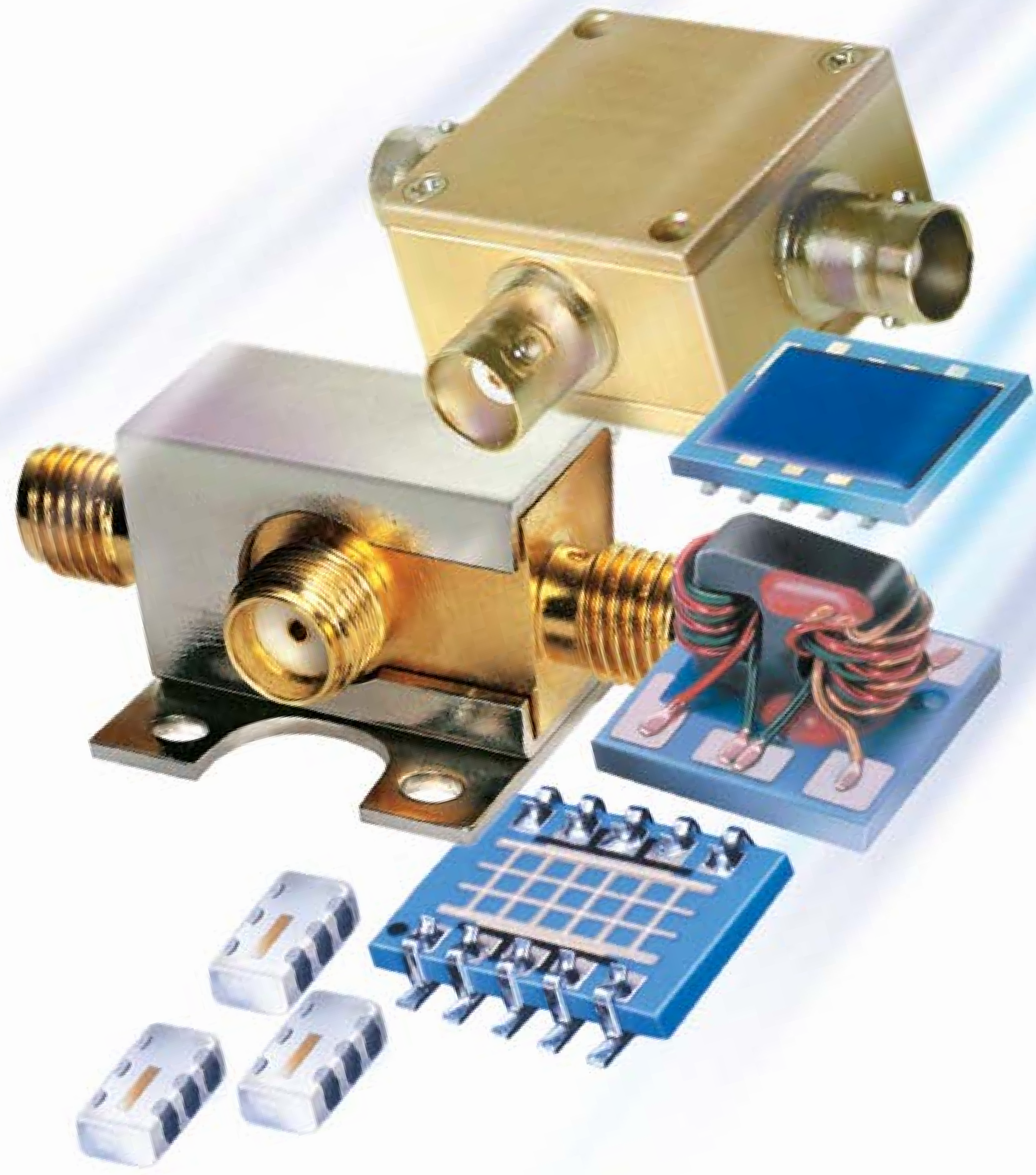

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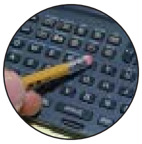

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www.rfic2008.org

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www.icuwb2008.org

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■ **Site:** Menlo Park, CA
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■ **Site:** Atlanta, GA
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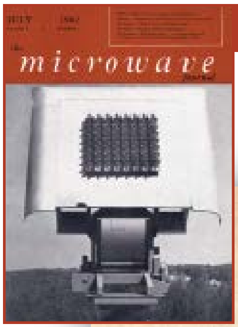

SIMILARITY CONSIDERATIONS FOR VARACTOR MULTIPLIERS*

DR. A. UHLIR, JR.
SEMICONDUCTOR RESEARCH & DEVELOPMENT
MICROWAVE ASSOCIATES, INC.
Burlington • Massachusetts

INTRODUCTION

This paper is concerned with the connection between the varactor equivalent circuit and harmonic generator performance. Substantial progress has recently been made in the detailed analysis of multiplier circuits by several workers; these results will likely be published in the near future. Instead of borrowing from these new results, the present paper will show the extent to which the problem can be surrounded by general arguments based mainly on dimensional analysis.

In fact, the combination of a few early experimental results with dimensional analysis provides a basis for preliminary multiplier design — particularly the initial selection of varactor diodes.

The discussion can and will be carried out without any circuit diagrams. But it will be apparent that important existence assumptions about circuits are involved. It is implicitly assumed, for example, that at least one optimum multiplier circuit exists for a given order of multiplication.

QUALITATIVE DESIGN CONSIDERATIONS

One of the most important qualitative considerations in multiplier and multiplier chain design is the harmonic number. If the input and output frequency are predetermined, the choice is restricted to the prime factors of the frequency ratio, or products thereof. When the requirement is solely for an output at a given frequency, the designer is at liberty to choose the input frequency. The last is the case, for example, in self-contained solid-state microwave power sources consisting of crystal-controlled transistor oscillators, transistor power amplification at some VHF frequency, and multiplication by a varactor chain.

In the recent history of varactor multipliers, some have expressed the belief that doubler chains would be vastly more efficient, for converting a low frequency to a much higher frequency, than any combination of higher order multipliers. This idea arose not only from experiments but also from approximate theoretical analyses.[1,2] This view cannot be confidently held today in the face of more recent experiments.

The key to obtaining good efficiencies in high order multipliers is the use of "idler" circuits which permit current flow at intermediate harmonics.[3] One must permit current to flow at the second harmonic to generate appreciable quantities of *any* higher harmonic; other idler frequencies are also needed for best results in multiplication by more than four times .

It is true that a doubler has many advantages. It is the simplest type of frequency multiplier. It is also the most efficient, but it cannot be compared to higher-order multipliers on the basis of efficiency alone, because the higher-order multipliers accomplish a larger frequency conversion.

An application where the doubler excels is in generation of the highest frequencies. When the output frequency is as much as one-tenth of the varactor cutoff frequency, the doubler is decidedly more efficient than any higher-order multipliers. Thus, with present varactors, any multiplier chain intended to generate millimeter-wave power should have a doubler as the final stage.

It seems obvious that doubler chains are to be preferred when a wide-band swept output is to be obtained. In principle, one might approach octave bandwidths.

However, there are severe difficulties in designing efficient multiplier chains, even using doublers, for wide input frequency ranges. Resonant circuits which function satisfactorily over one or two per cent bandwidth have to be

* *Supported in part by the Electronic Technology Laboratory of the Wright Air Development Division, under Contract No. AF33(616)-7232.*

replaced with more complicated broadband filters. To accomplish this broadbanding without incurring parametrically excited oscillations at some points of the band, one is usually forced to sacrifice efficiency.

Higher-order multipliers are sometimes suggested by frequency allocations or existing low-frequency equipment. For example, triplers are popular for adapting 150 Mc mobile radio equipment to 450 Mc bands, 2 kMc microwave relay components to 6 kMc, and 3 kMc radar components to 9 kMc.

BIAS

Most of the available experience on high-power, high-efficiency multipliers is with silicon varactors. Semiconductor materials such as germanium and gallium arsenide will differ primarily in two factors: dc bias and in the limitations on capacitance variation imposed by the conductance associated with the junction.

To the extent that the device is a variable capacitor rather than a rectifying element, the dc bias should be specified in terms of voltage rather than current. The suitability of operation without external bias might be expected to depend upon the breakdown voltage of the junction and the power level of operation. High efficiency multipliers have been made with open-circuit dc for silicon varactors having breakdown voltages from 6 to 120 volts.

One of the advantages of operating without an external bias is that bias leads are subject to pickup. Multipliers are quite sensitive to pickup, since they act as up-converters and amplify low-frequency interference in the process of converting it to sidebands of the harmonics. Mercury cells may be built into the multiplier to avoid the pickup and filtering problem if it is found that bias is desirable.

Some varactor multipliers have been operated with an optimized dc return resistance. Such circuits require that the rectification characteristics of the varactor be controlled.

VARACTOR SELECTION

The requirements on the varactor diode for power harmonic generation include, first, the ability to convert the given power without being driven beyond its voltage limitation. This property will be termed "reactive power capability" and is characteristic of the junction element.

The second problem is to have low electrical losses for efficiency and minimum heat generation. The series resistance, as represented by the cutoff frequency, has generally been the dominant loss mechanism, with rectification secondary.

The third requirement is to have a structure with sufficient thermal dissipation to carry away the heat generated by the electrical losses remaining after everything possible has been done to minimize these losses.

The distinction between thermal dissipation and reactive power capability can be illustrated by considering short pulses of fundamental power. For very short pulses (including, nevertheless, many cycles), the power dissipation can be reduced to a small value, but the varactor will not function properly unless it has sufficient reactive power capability for the peak power.

SIMPLE ESTIMATES OF REACTIVE POWER CAPABILITY

For specified capacitance-voltage characteristics, it should in principle be possible to analyze analytically or with a computer the detailed performance of any harmonic generator circuit.

However, very simple equations can be used for practical selection of varactors. These relationships can be derived by similitude (dimensional analysis). Similitude is easily applicable to varactors because all of the leading types can be approximately represented by a capacitance-voltage characteristic of the following form:

$$C(V) = \frac{C(0)}{(-V+\phi)^{\gamma}} \qquad -V_B < V < +\phi \qquad (1)$$

where V is the voltage, ϕ is a so-called "built-in" potential, and γ is an exponent that usually lies in the range of 0.33 to 0.50. This relation applies to the depletion-layer capacitance. Other mechanisms carry one outside of the present theory, as discussed below under LIMITATIONS. Equation (1) may be rewritten in the form

$$C(V) = C_{min}\left(\frac{V_B+\phi}{-V+\phi}\right)^{\gamma} \qquad (2)$$

where $C_{min} = C(-V_B)$. This form of the relation has two advantages. One of passing interest here is that the temperature dependence can be almost completely represented by temperature variation of the parameter ϕ in this Equation (2), but not in Equation (1) because of variation in C(0).

The particular significance of Equation (2) for the present dicussion is that *two* quantities, $\phi + V_B$ and C_{min}, characterize the relation apart from the choice of the dc bias voltage. In everything that is said further, it will be assumed that the dc bias is adjusted to the optimum value. Obviously, a theory of this nature cannot give logical assurance that operation without dc bias will be successful. Experiments alone afford encouragement in this direction.

For present diodes, V_B is at least several volts, while ϕ is less than one volt. Thus, V_B can be used as a good approximation for $\phi + V_B$ in all calculations.

The reactive power formula can be derived by a simple argument without formal dimensional analysis. Current through a capacitance is proportional to the capacitance, to the frequency, and to the rf voltage. Under conditions of similarity, the rf voltage will be proportional to $\phi + V_B$ or, approximately, V_B. Hence, the current i will be proportional to f C_{min} V_B. The volt-amperes or reactive power will therefore be given by

$$VA = \frac{K}{2} f_{in} C_{min} V_B^2 \qquad (3)$$

In this equation, K is a dimensionless quantity depending on n and f/f_c, for a given C —V characteristic. The denominator 2 has been included as an arbitrary constant. One can remember the quantity $C_{min} V_B^2/2$ as the energy stored at a voltage V_B in a linear capacitor having a capacitance C_{min}. The amount of energy that can be stored in the varactor is proportional to this quantity, as is the power that can be transformed per cycle. The use of input fre-

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APA3438-35	3.4 to 3.6	30	3.5	0.5	35	45	2.8 A
APA3438-38	3.4 to 3.6	30	3.5	0.5	38	48	3.4 A
APA3438-41	3.4 to 3.6	30	3.5	0.5	41	51	5.5 A
APA3642-39	3.6 to 4.2	40	4.0	1.0	39	49	3.3 A
APA4450-40	4.4 to 5.0	40	4.0	1.0	40	50	5.5 A
APA4450-42	4.4 to 5.0	40	4.0	1.0	42	52	3.3 A
APA4450-44	4.4 to 5.0	40	4.0	1.0	44	54	9.5 A
APA5964-36	5.9 to 6.4	40	4.5	0.5	36	46	2.6 A
APA5964-42	5.9 to 6.4	40	4.5	1.0	42	52	5.3 A
APA5964-44	5.9 to 6.4	40	4.5	1.0	44	54	9.5 A
APA5864-46	5.8 to 6.4	40	4.5	1.0	46	56	11.0 A
APA6472-36	6.4 to 7.2	40	4.5	0.5	36	46	2.6 A
APA6472-42	6.4 to 7.2	40	4.5	1.0	42	52	5.9 A
APA7785-39	7.7 to 8.5	40	4.5	1.0	39	49	3.6 A
APA1112-36	10.7 to 11.7	40	4.5	0.5	36	46	2.6 A
APA1112-42	10.7 to 11.7	40	4.5	1.0	42	52	5.9 A
APA1414-37	14.0 to 14.5	40	4.5	0.5	37	47	2.6 A
APA1414-40	14.0 to 14.5	40	4.5	0.5	37	47	2.6 A
APA1414-43	14.0 to 14.5	40	4.5	0.5	37	47	2.6 A
APA3031-36	30.0 to 31.5	27	6.0	1.0	36	43	5.6 A

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quency f_{in} is an arbitrary choice from the standpoint of dimensional analysis but seems to make K less dependent upon harmonic number n than would be the case if output frequency were used. Values of K up to 10 have been observed for efficient VHF multipliers, but K = 0.5 to 1.0 seems to be a conservative range to use in initial design, especially in the microwave region. The wide variations in observed values of K suggest that the basic equivalent circuit is an oversimplification.

At present, no serious miscalculations would result from the notion that the varactor must store a certain amount of energy and that it delivers this energy to a harmonic once per cycle. However, it may be important to disavow this implication to preclude possible future misunderstandings. We do not ask the varactor to *store* energy — it must *transform* it from one frequency to another. Storage and transformation are not the same thing. They are both, of course, proportional to junction area and this is what Equation (3) basically indicates. For a given type of junction material, with a fixed breakdown voltage and capacitance-voltage characteristic, the ability to transform energy is proportional to area and for this reason proportional to capacitance.

Similarly, it would be a potential mistake for the future to suppose that a large value of C_{min} is desirable. This is not true out of the context of an assumed capacitance-voltage characteristic. On the contrary, it would be entirely desirable if a fixed amount could be subtracted from the varactor capacitance for all voltages.

THE EFFECT OF CUTOFF FREQUENCY

Similitude between different multipliers can be exact only for equal values of the ratio of the operating frequency to the cutoff frequency $f_c = 1/2\pi R_s C_{min}$. The cutoff frequency is a convenient way of bringing the series resistance into dimensional analysis of multipliers; thus

$$\eta = \eta\,(n, f_{out}/f_c) \tag{4}$$

where f_{out} is output frequency.

One can make a simple argument that goes beyond dimensional analysis, as follows.

The effect on the efficiency of a small series resistance R_s will be an i^2R_s loss at all frequencies at which currents flow, (This includes of course the input frequency and the output frequency but generally currents at other frequencies, especially for high-order multipliers). If one assumes that the currents themselves are not affected to first order by the small series resistance, the efficiency η must have the approximate form

$$\eta \approx \eta_o - \text{const. } R_s \quad (R_s \text{ small}) \tag{5}$$

where η_o is the efficiency in the absence of series resistance ($\eta_o \approx 1$ unless special restrictions are placed on the type of circuit). By combining Equations (4) and (5), one obtains the result for "efficient" multipliers

$$\eta \approx \eta_o \left[1 - \alpha_n\,(f_{out}/f_c) \right] \tag{6}$$

where α_n is a dimensionless constant.

An equally good way of writing the approximation is

$$\eta \approx \eta_o e^{-\alpha_n\,(f_{out}/f_c)} \tag{7}$$

which is equivalent to Equation (6) for small values of R_s. However, Equation (7) is found empirically to have a wider range of validity than Equation (6); Equation (7) is probably useful for efficiencies as low as 10 per cent. Also, Equation (7) is especially convenient for discussing multiplier chains (see below).

For doublers, first estimates of α_2 indicate a value of about 10 for the entire range of $\gamma = 0.33$ to $\gamma = 0.5$. For the triplers, it appears that α_3 is about 15. Both of these values are provisional and subject to more accurate determination. It may eventually be established exactly how they vary as γ varies. The difference between η_o and unity represents the generation of unwanted harmonics and reflection of the fundamental.

MULTIPLIER CHAINS

Since the overall efficiency η_T of a complete multiplier chain is the product of the efficiencies of the individual stages, the approximate formula for efficiency, Equation (7), is an especially convenient form for calculating the overall efficiency. In particular, one is often interested in the best choice of harmonic orders to construct a chain. This question will be treated for the special case where all varactors in the chain are assumed to have the same cutoff frequency and each multiplier is of the same order n. Then, using m multipliers to generate a frequency n^m times the input frequency f_{in}, one has

$$\eta_T = e^{-(n\alpha_n f_{in}/f_c \,+\, n^2\alpha_n f_{in}/f_c \,+\, \ldots\, n^m\alpha_n f_{in}/f_c)} \tag{8}$$

Expressing the sum in closed form, one has

$$\eta_T = e^{-\frac{n}{n-1}\alpha_n \frac{f_{in}}{f_c}(n^m - 1)} \tag{9}$$

or

$$\eta_T = e - \frac{n}{n-1}\alpha_n \frac{f_{out} - f_{in}}{f_c} \tag{10}$$

One can see that the quantity $\frac{n}{n-1}\alpha_n$ is a measure for comparing the various orders of multiplication. On this basis, doubler and tripler chains are practically equivalent; *i.e.*, the choice of order of multiplication will not be made on the basis of theoretical efficiencies.*

---

* *A theoretical analysis of multipliers (with idlers where required) has been made by B. L. Diamond, Lincoln Laboratory report 476-0012, 13 December 1960, using the depletion-layer capacitance assumption with $\gamma = 0.5$ (abrupt junction type). Examination of his results (by graphical methods, hence approximate) gives for doublers, triplers, quadruplers, and sextuplers, repectively, the following values for α_n: 10.6, 13.0, 16.6, 18.8, and for the quantity $\frac{n}{n-1}\alpha_n$: 21.2, 19.5, 22.1, 22.5. The values of α_2 and α_3 are quite close to the cited early experimental results for varactors of the graded junction variety ($\gamma = 0.33$). The near-equality of the values of $\frac{n}{n-1}\alpha_n$ shows that the theoretical efficiency of a chain does not depend much upon order of multiplication. It is expected that this analysis will be published.*

It is interesting to note that the last stage of such a (theoretical) chain always has a lower efficiency than the combined efficiency of all previous stages, no matter how low the starting frequency. It also follows that the optimum fundamental frequency will not be determined on the basis of theoretical efficiencies, but on considerations such as simplicity and signal purity.

The above conclusions are based on the approximations for "efficient" multipliers. When the required output frequency is so high that the last stage must be inefficient, the last stage should be a doubler regardless of the arrangements used in the lower-frequency parts of the chain.

Also, the simple relations like Equation (10) are based on the assumption of constant cutoff frequency, while the reduced reactive power requirements at the high-frequency end of a chain permit one to choose varactors of the highest cutoff frequencies.

IMPEDANCE CONSIDERATIONS

For a given reactive power, a variety of varactors may serve having different capacitances and breakdown voltages but the same nominal reactive power. The input and output admittances are proportional to C_{min} and whenever high power and high frequency are involved, increase in power by use of larger values of C_{min} leads to very low circuit impedances. For this reason, in multiplier stages below 1000 Mc output frequency, it is generally desirable to use a relatively high breakdown voltage.

A somewhat more quantitative estimate of impedance may be suggested at least for doublers and triplers. In the low-frequency region where high efficiencies are obtainable, one can reasonably expect the reactance corresponding to C_{min} to be at least 10 times the source impedance. The reactance of C_{min} at the output frequency would then be three to five times the output impedance if the latter were equal to the source impedance. Of course, any impedance estimates based on the diode characteristics can be modified by the use of transformers in the input and output circuits. Such transformation should be particularly easy to achieve for narrow-band multipliers. Accordingly, while a 50 ohm source impedance is the commonest preference in UHF and microwave circuitry, it is not unreasonable to consider a varactor that would suggest a five ohm or even lower source impedance. On the other hand, a 500 ohm source impedance seems to be a reasonable upper limit. Therefore, the capacitive reactance of C_{min} should fall in the range of 50 to 5000 ohms with a preference for 500 ohms.

At the other extreme, where the cutoff frequency is not much higher than the signal and output frequencies, the effective impedance of the varactor must approach the series resistance. Optimum power transfer will thus require source and load impedance approximately equal to the series resistance as transformed to the varactor. Such multipliers are very inefficient, of course.

The cutoff frequency of the lower-voltage varactors is generally better than that of high-voltage varactors (although the margin of difference is not as large as formerly). For any type of junction, however, it is difficult to make (or to measure?) series resistances much below 0.5 ohms.

As frequency is increased, it becomes desirable to use low-voltage varactors in low impedance circuitry to take advantage of high cutoff frequency (often 200 kMc). Skin-effect losses increase with frequency, making it more difficult to work at low impedances at high frequencies. This dilemma is no surprise to anyone experienced in microwave circuits; difficulties usually increase when frequency goes up. Each case must be examined in detail, but the trend is toward the use of lower-voltage varactors at higher frequencies and the total power output per varactor accordingly diminishes with frequency.

NORMALIZATION POWER VS. REACTIVE POWER

The cutoff frequency and the nominal reactive power tell a great deal of information about the electrical characteristics of the junction. An alternative involves the use of cutoff frequency and "normalization power" P_n.[4] As far as information content is concerned, these two sets of parameters are equivalent since they are related by the formula

$$P_n = 4\pi f_c P_r = (V_B + \phi)^2/R_s \qquad (6)$$

The nominal reactive power P_r is preferable if one is concerned with comparatively efficient multipliers and wishes to specify varactors that are likely to be replaceable in a given circuit. The normalization power approach has some advantage if one wants to estimate maximum power that can be extracted at a given frequency from a particular varactor by optimizing the circuit especially for the chosen diode.

Suppose, for example, that an efficient multiplier circuit was designed for a given varactor and then improvement in fabrication techniques resulted in the possibility of reducing the series resistance without changing any of the other characteristics of the junction. The improved varactor would have to have the same capacitance and nominal reactive power as the older type. A higher efficiency (and better reliability because of cooler operation) would result from the improved cutoff frequency.

Thus, for efficient multipliers, one should think of the series resistance as a second-order parameter. For this reason, the nominal reactive power characterization seems preferable to the normalization power, in which the series resistance appears as a first-order quantity.

LIMITATIONS

While the above design theory is a logical consequence of the assumptions, the assumptions are not completely faithful to the physics of the device. For example, the assumed capacitance-voltage characteristic implies that the capacitance goes to infinity at a forward voltage equal to ϕ. At this voltage, the hole and electron clouds on either side of the junction are supposed to meet each other. Of course, such an abrupt event does not occur because of the thermal fuzzing of the exact boundaries of the distribution. Moreover, the equations imply that the amount of charge introduced into the diode in the forward direction is finite; at least the equations do not say what should happen after this point is reached. In fact, the holes and electrons continue to approach each other and intermingle. Some of this intermingling occurs even before the depletion region is obliterated and it continues as a charge-storage mechanism for an almost unlimited amount of charge.

This mechanism is useful only at sufficiently low frequencies where the delay due to the finite speed of carrier motion is a very small fraction of a cycle. The delay is, of course, dependent upon the junction structure and is expected to be shorter for diodes with lower breakdown voltage (because the junctions are narrower).

Charge-storage by the intermingling of holes and electrons (sometimes called "minority carrier storage" or "injection capacitance") would be expected to increase the power handling capabilities almost without limit, according to the following reasoning. This mechanism leads to a capacitance-voltage characteristic of the form[5]

$$C \approx \text{constant} \times e^{qV/kT} \qquad (11)$$

This capacitance becomes dominant for forward biases. At comparatively low frequencies, the reactance of this large capacitance may be much larger than the series resistance and therefore useful in harmonic generation. In fact, there can be a wide range over which R_s and C_{min} are electrically insignificant and the predominant characteristic is the capacitance given by Equation (11). In this range one has the remarkable result that the reactive power capability is not dependent upon junction area, since one can increase the capacitance as desired by appropriate change in voltage. Also, as the power is increased for a given diode, the effective capacitance (however defined) will increase, the operating cutoff frequency will decrease (if R_s is approximately constant), and the efficiency will decrease. This behavior differs from the depletion-layer case where efficiency increases with power level up to the reactive power limit.

All of the predicted phenomena have been observed in high power VHF multipliers: anomalously high values of 10 to 20 for K in Equation (3), reduced dependence of electrical performance upon junction area, and gradual decrease of efficiency with increasing power.

Larger-area junctions have the advantage of lower R_s and better thermal dissipation, so that ultimate power capability is approximately proportional to area. The ultimate reactive power capability is so enhanced by the injection capacitance that the i^2R_s thermal limitation is usually the controlling factor in determining maximum CW power.

Multiplier testing of power varactors.

CONCLUSIONS

The systemization of varactor multiplier results by the methods of dimensional analysis shows that additional experimental work is required to develop a better equivalent circuit to represent the varactor in the VHF region. Detailed mathematical analyses based on the depletion-layer will nevertheless be of considerable value, since it is likely that this equivalent circuit is approached in the microwave region.

The results that have been derived for efficient multipliers show that theoretical analyses of efficiency will not likely form the basis of choosing the harmonic order to be preferred in practical multipliers, nor will such theories provide an indication of optimum fundamental frequency. These choices will be based on practical considerations such as circuit simplicity and transistor availability.

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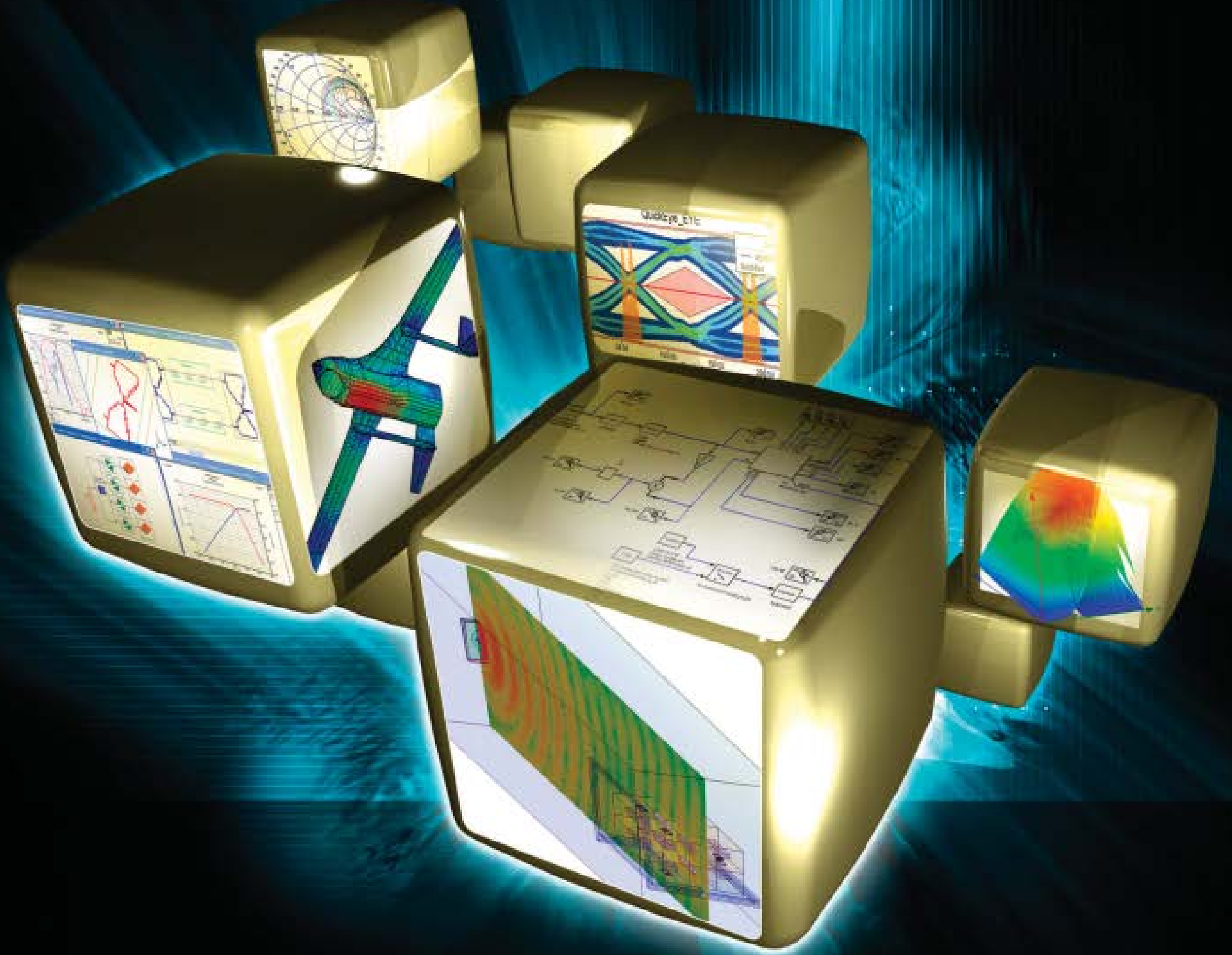

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AN HISTORICAL PERSPECTIVE ON 50 YEARS OF FREQUENCY SOURCES

This year is the 50th anniversary of *Microwave Journal*. As is frequently the response to such noteworthy occasions, the editors, in their unquestionable wisdom, have invited old codgers like me to create retrospectives on microwave technology. I never thought it would have happened, but the older I get, the more I tend to say things like “in my day” and to mutter more or less conventional complaints about the younger generation. In technology, we happily become the old guys that, in our youth, so thoroughly annoyed us.

This month, *Microwave Journal* is highlighting a 1962 article by Arthur Uhlir, one of the true pioneers of nonlinear component research. The article, while seemingly rather specialized, touches on many of the then-current considerations in the design of such components. In view of all this, it seems like a good occasion to talk a little about sources, and to show how we got where we are.

VACUUM SOURCES

At the time of Uhlir’s paper, there were few options for generating microwave signals. The simplest, although rarely the most desirable, was a microwave tube.[1,2] Three types of tubes used for generation of microwave signals were magnetrons, reflex klystrons and backward-wave tubes. More or less conventional triode tubes could also be configured to work in the lower microwave range.

Magnetrons

Magnetrons, invented in England during the second world war, were the earliest microwave tubes. Magnetrons, more than any other technology, made radar possible, and, because of their low cost and high power ca-

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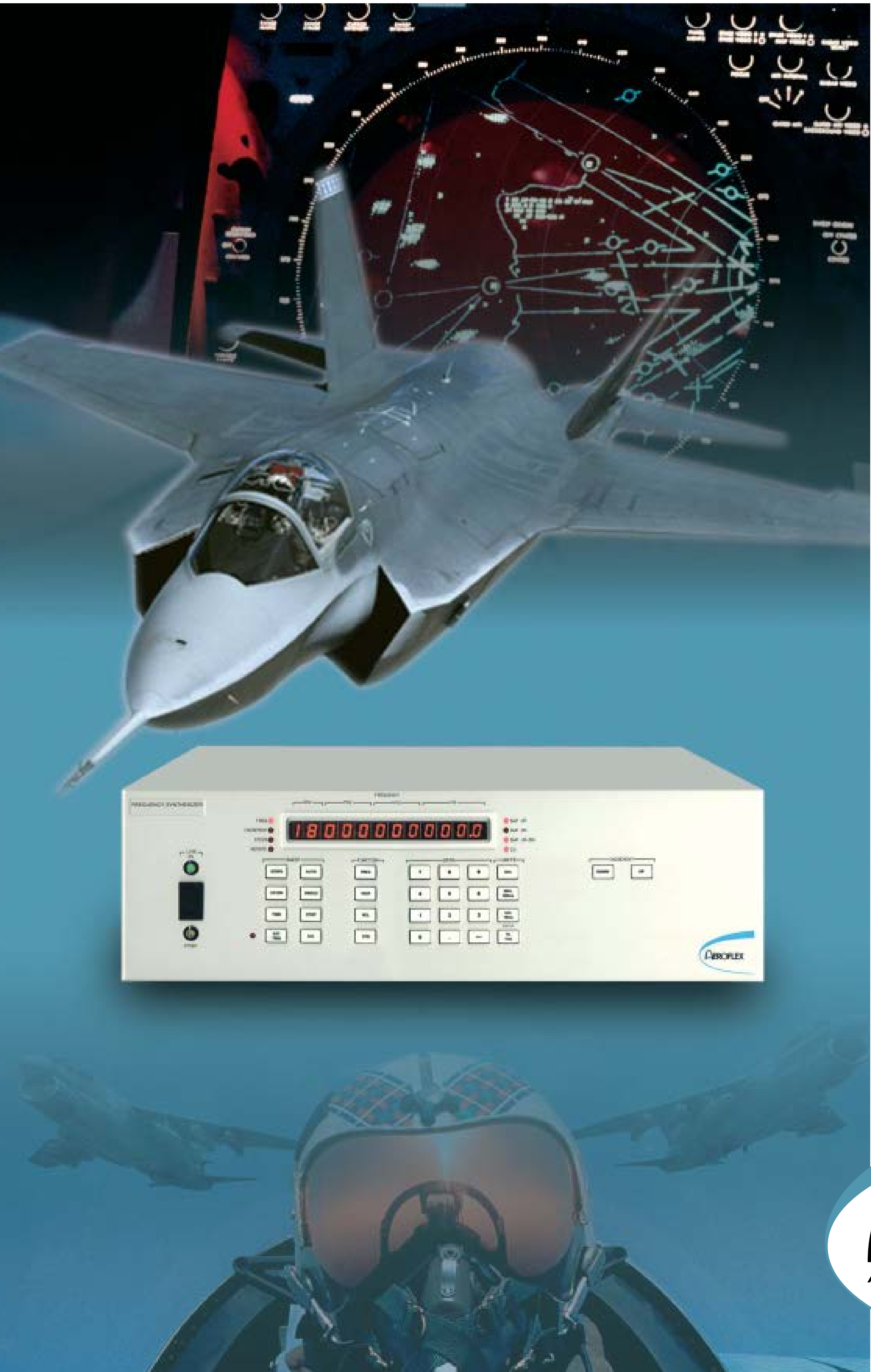

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pability, are still in use. Your microwave oven, for example, uses one.

Magnetrons operate by accelerating electrons and spiraling them in a magnetic field. The electrons pass apertures in a periodic structure composed of resonators, giving up energy to the electric field in the aperture (see ***Figure 1***). The principle has been compared to creating a tone by blowing across the open top of a bottle. Like other microwave tubes, magnetrons are noisy and difficult to stabilize. As such, they are rarely used for receiver local oscillators, but for applications that require large amounts of raw power, such as radar transmitters and, of course, your microwave oven, they are ideal.

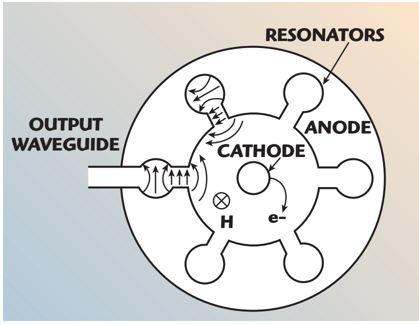

Fig. 1 A magnetron is one of the simplest and earliest microwave tubes. Electrons emitted from the cathode are accelerated toward the anode, but a magnetic field causes them to spiral in the space between those elements. The spiraling electron cloud excites fields in the coupled cylindrical cavities in the anode.

Reflex Klystrons

Reflex klystrons were common at frequencies from the lower microwave range well into the millimeter. In an amplifying klystron (see ***Figure 2***), an electron beam first travels through a microwave cavity, where it is velocity-modulated by the electric field. The beam continues through a drift space, where the faster electrons catch up to the slower ones, forming "bunches" of electrons in the beam. When the "bunched" beam finally passes through a second cavity, it gives up some of its energy exciting the fields in that cavity, providing greater output power than input power; that power, in effect, is extracted from the beam. In a reflex klystron, only a single cavity is used. A reflector element bends the beam back into that same cavity that velocity-modulated it, effectively providing feedback necessary for oscillation. Reflex klystrons were capable of providing high power, but they were difficult to stabilize and generated substantial AM noise. In many millimeter-wave receivers, the AM noise from the klystron, downconverted by the klystron local oscillator (LO) itself, was the dominant source of noise.

Backward-wave Tubes

A backward-wave oscillator (BWO) is a variant of a travelling wave tube, or TWT. In a TWT, a wave is allowed to interact with an electron beam over a fairly long region (see ***Figure 3***). The wave must propagate on a slow-wave structure, usually a helix, so its velocity matches that of the beam. Under the right conditions, the wave modulates the beam and the beam gives up energy to the wave, in a manner not terribly different from a klystron. The main difference, however, is that a cavity is not used, so TWTs can operate over broad bandwidths.

Backward-wave oscillators are possible because, contrary to all intuition, the beam can support a backward wave resulting from the forward-travelling wave. This creates the feedback necessary for oscillation. Furthermore, just as TWTs are broadband, BWOs can have remarkable tuning ranges. Many of us remember using BWO test sources, from the lower microwave region to the millimeter-wave, having octave-band outputs.

While BWOs are largely obsolete, TWTs are still in use. Although the preference today is for solid-state amplifiers, many spacecraft in use today still have TWTs. They are used because of their high efficiency, high power, reasonable linearity and broad bandwidth. Especially, TWTs can be made adequately reliable for space applications. It's not an exaggeration to say that, without TWTs, broadband, high-volume satellite communication would probably not have been possible until very recently.

Although they were essential for early microwave technology, vacuum devices had limitations that encouraged system designers to abandon them as soon as adequate solid-state alternatives were available. As we have noted, tubes were notoriously difficult to stabilize. Throughout the 1960s and 70s, communication systems became progressively more sophisticated, and frequency and phase stability requirements became more severe. Although it is possible to phase-lock certain kinds of tubes (the reflex klystron is probably the easi-

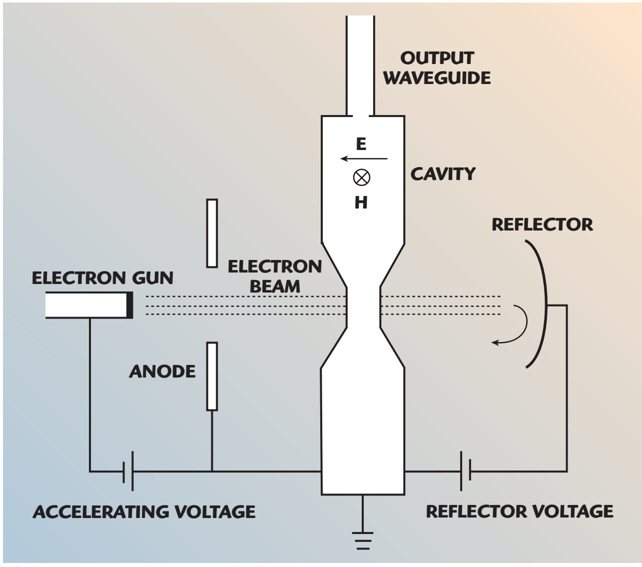

Fig. 2 In a reflex klystron, the beam is first velocity-modulated in a microwave cavity, then reflected back into the cavity to provide feedback necessary for oscillation. As the cavity is part of the body of the tube, it is necessarily at ground potential and a negative voltage is applied to the cathode.

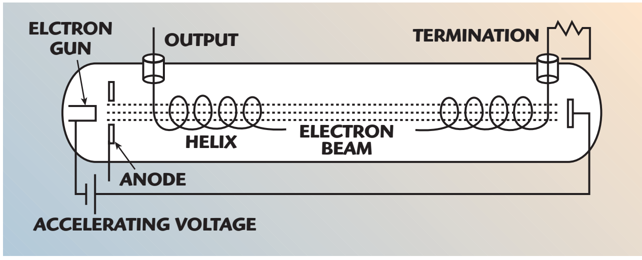

Fig. 3 A backward-wave oscillator is similar in many respects to a travelling-wave tube, but it makes use of a backward wave that propagates on the helical slow-wave structure. This provides feedback necessary for oscillation.

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S6W2	S6W5	N6W5	6	±0.40
S7W2	S7W5	N7W5	7	-0.4, +0.9
S8W2	S8W5	N8W5	8	±0.60
S9W2	S9W5	N9W5	9	-0.4, +0.8
S10W2	S10W5	N10W5	10	±0.60
S12W2	S12W5	N12W5	12	±0.60
S15W2	S15W5	N15W5	15	±0.60
S20W2	S20W5	N20W5	20	-0.5, +0.8
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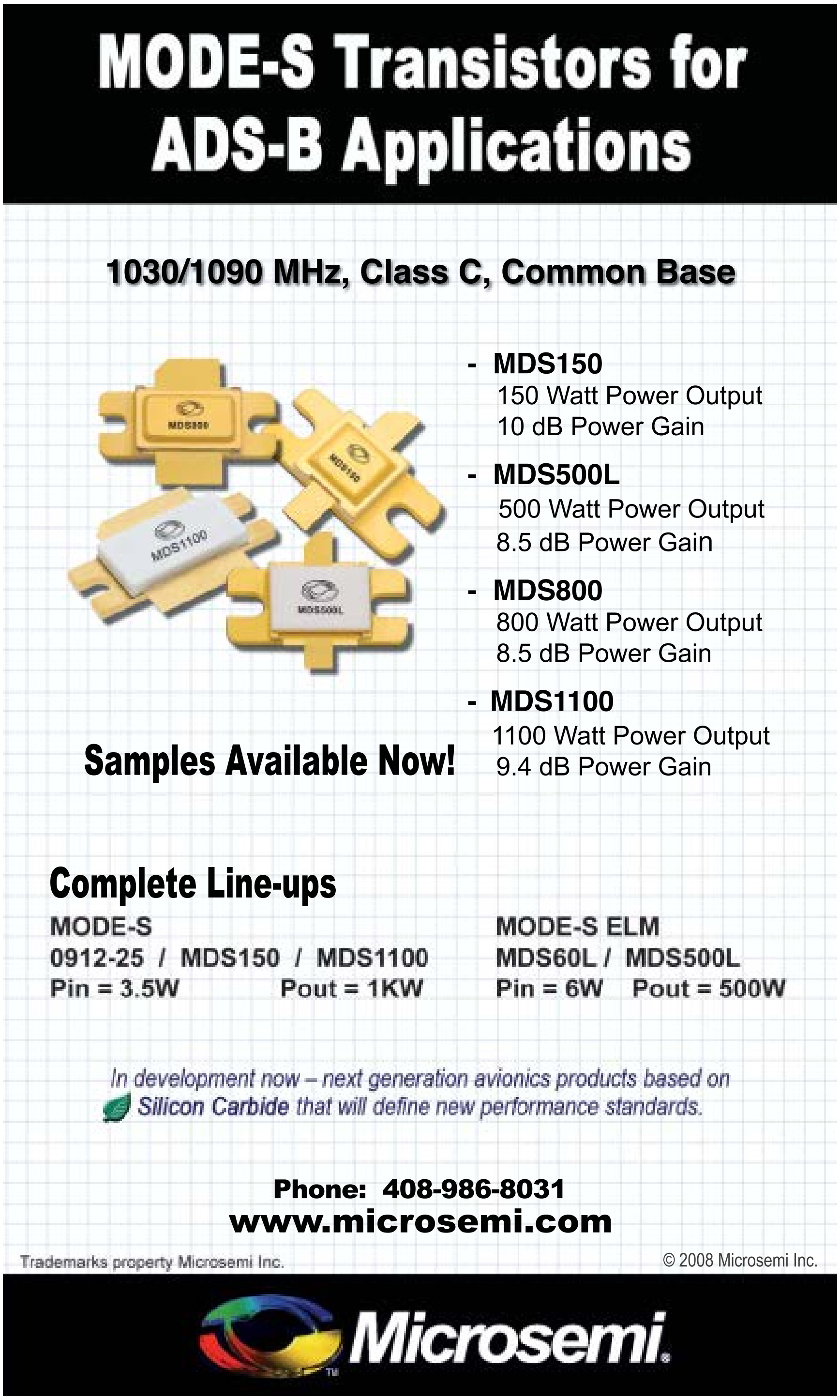
RF/IF MICROWAVE COMPONENTS 331 Rev K est), it is indeed not much fun to try. Even then, the AM noise problem was severe, especially for low-noise millimeter-wave receivers used in radio astronomy and other radiometric applications. Finally, and probably most importantly, the problem of reliability motivated the move from tubes to solid-state sources. All tubes have a lifetime, usually no more than a few thousand hours of operation. Even so, and perhaps paradoxically, the tubes themselves often were reliable enough, but the necessary high-voltage power supplies were not. High voltage is inherently unreliable, especially in space applications, where rarefied gasses are easily ionized and support arcing. Another motivation was safety; as this author, who once superimposed himself across the 1800 V output of a klystron power supply can attest, high voltages are inherently dangerous.



Conventional Vacuum Tubes

Even before the 1960s, a number of vacuum tubes could generate useful signals at UHF and lower microwave frequencies.[3] Most of these were triodes, tubes having a single grid. World War II military equipment often used the 955 "acorn" tube, invented in 1935.[4] I have seen 955s used up to 500 MHz, but with proper cavity construction, they could possibly operate to 1 GHz. The 2C40 "lighthouse" tube's cylindrical symmetry made it ideal for mounting in coaxial cavities. 2C40s could oscillate above 3 GHz. The 5675 "pencil" tube, also designed for use in coaxial cavities, could produce a half watt at 1.7 GHz. Finally, one of the best known tubes, and perhaps the last gasp of high-frequency receiving-tube technology, was the 6CW4 "Nuvistor" from RCA.[5] It was used in many high-frequency commercial applications, primarily television tuners, and could provide low-noise oscillation and amplification to almost 1 GHz.

In the solid-state world, we focus almost entirely on resistive and capacitive parasitics. Conventional vacuum triodes, however, were limited in their operation by cathode-to-plate electron transit times. Making devices smaller and operating at higher plate voltage reduced the transit time; unfortunately, small size and high voltage were competing trade-offs and, in any case, a triode could be made only so small. It seems inevitable, then, that transistors, which have much shorter transit times, would be viewed as the next high-frequency devices, and it is no surprise that great effort was put into the development of microwave transistors. Indeed, not only were transistors improved beyond all expectations, but new, unforeseen solid-state microwave devices were created as well. It's not an exaggeration to say that solid-state technology created a golden age of microwave electronic technology.

SOLID-STATE SOURCES

Varactor Frequency Multipliers

At the time of Uhlir's paper, practical microwave transistors didn't exist. Rapid progress in solid-state technology was made throughout the 1960s, however, and by the early 70s we had good, low-noise bipolar transistors useful to a few gigahertz.

The earliest solid-state device useful for generating microwave energy was the varactor diode. Varactor frequency multipliers date from the late 1950s, when they were realized in both diffused-junction and point-contact form. It seems that good varactors, useful for efficient microwave frequency multiplication, were available quite early, and practical multipliers were produced well before 1960. While some aspects of the operation of reactive microwave devices were becoming clear by the mid-1950s (for example, the classic treatment of power in such devices by Manley and Rowe),[6] a good understanding of the theory of such devices followed well behind practical application. For example, another paper by Uhlir at this time[7] shows that the operation of even simple point-contact diodes (in which a metal contacting "whisker" actually forms a crude Schottky junction) was not well understood.

A good theoretical understanding of varactor multipliers was developed fairly quickly, however, and papers of that era show a remarkably complete theoretical basis for varactor multipliers.[8,9] The paper by Uhlir, reprinted in this issue, shows a mature understanding of multiplier theory by 1962. One of my favorite books on varactor circuits, although focussed primarily on parametric amplifiers, is that of Blackwell and Kotzebue.[10] It hit the streets in 1961. Not much later, Burkhardt's classic paper on multiplier design was published;[11] it became the standard for the design of such multipliers.

Early on, it was recognized that a varactor diode's voltage was a quadratic function of charge, and, as such, could generate only second harmonics directly. Higher harmonics required the use of idlers, resonators that supported voltage components at intermediate harmonics, allowing higher harmonics to be generated through a mixing process. Although Manley and Rowe showed that 100 percent efficiency was theoretically possible in reactive multipliers, the efficiency, in practice, decreased fairly quickly with harmonic number. Uhlir's paper takes issue with that idea, largely for theoretical reasons, and it is important to note that it was published before more complete analyses such as Burkhardt's. Even so, it indicates quite a bit of activity in the theoretical work surrounding these components.

In the early 1960s, there existed few solid-state methods for generating microwave signals. The most common was to use a crystal oscillator, operating below 100 MHz, followed by a series of bipolar-transistor frequency multipliers. The multipliers raised the frequency to a few hundred megahertz, at which the signal was amplified and applied to a string of varactor multipliers. If the multiplier chain had more than a few varactor stages, the input power to the varactor portion often had to be a few watts. This would result in a few tens or hundreds of milliwatts at the output. To attain this modest output power, the source had to dissipate several watts of heat. Not only was this inefficient, but disposing of the heat, especially in space applications, could be a major difficulty. Apart from efficiency considerations, varactor multipliers were generally narrowband and prone to instability.

Making them work well was a tricky business.

On the positive side, the use of a crystal oscillator guaranteed good stability. Transistor frequency multipliers generated little noise, and varactors, being reactive devices, generated only negligible levels of noise. As a result, such sources could be quite low-noise. This combination of low noise and stability made varactor multiplier chains extremely valuable in modern radar and communication systems.

Other Solid-state Sources

Probably the most successful solid-state microwave source is the Gunn oscillator, also known as the transferred-electron oscillator. Such oscillators make use of negative resistance that can occur in certain bulk semiconductors, all of which are III-V elements. These semiconductors have a "satellite" conduction band, in which the electron mobility is relatively low. When the electric field in the semiconductor is great enough, electrons are transferred into this band, their velocity decreases and current decreases. The resulting negative resistance can be used to create an oscillator.

The possibility of such negative-resistance oscillation was predicted in papers by Hilsum[12] and by Ridley and Watkins[13] before it was observed experimentally by Gunn.[14] Commercial oscillators were readily available by the early 1970s, and the technology quickly matured. Indeed, microwave engineers could hardly be luckier: put in DC, get out microwaves. What could be better?

Gunn oscillators had low noise, were easy to tune electronically (so they could be stabilized in phase-locked loops), could operate at millimeter wavelengths, and had decent output power and efficiency (a couple hundred milliwatts at 30 GHz was typical). GaAs Gunns had an upper frequency limit around 100 GHz established by the time required to transfer electrons to the satellite band. Other materials, particularly indium phosphide, were not so limited, and InP oscillators above 100 GHz were regularly produced. Not cheaply, of course, but for a price, they could be obtained.

The main disadvantages of Gunn oscillators were relatively high 1/f noise, which resulted in phase noise, the need to mount the device in a microwave cavity, and like all negative-resistance devices, a tricky design and development process. As a result, Gunn oscillators required a significant amount of labor in the form of fabrication and tuning to produce them. This meant, in turn, that costs were difficult to minimize, even with significant technological advances: making a Gunn oscillator still required several hours work by both a skilled machinist and an experienced technician. This probably was the greatest cause of the Gunn's eventual obsolescence.

Another important device was the impact-avalanche transit-time device, or IMPATT, along with its alphabet soup of cousins (TRAPATT, BARITT,…). IMPATTs are also negative-resistance devices, but their negative-resistance results from transit time through the device. The device, a kind of PIN diode, is allowed to go into avalanche breakdown at reverse-voltage peaks. Then, delays in the build-up of avalanche charge and transit through the device result in a current pulse that is out of phase with the voltage that generated it. The result is negative resistance and resulting oscillation.

Avalanche breakdown is a noisy process, so IMPATTs were unavoidably noisy. They were rarely used in receivers. On the positive side, however, they were capable of substantial power, often tens of watts at X-band. This made them serious contenders for tube replacement in various kinds of transmitting hardware. Even though the thought of using a device based on avalanche breakdown sent reliability engineers into post-traumatic shock, IMPATTs were eventually made reliable enough for space applications. They replaced a lot of TWTs in microwave systems.

WHAT DOES ALL THIS TELL US?

Of course, it would be easy to repeat the tired idea that those who don't learn from history are condemned forever to quote Santayana (or something like that). Even so, I find that there is often little understanding among technologists of the history of their technologies, and, especially, the phenomena that motivated their evolution. A result is research into technologies (often in academic settings, I'm afraid) that seem to push that evolutionary process backwards. The best way to prevent that malady is to understand that evolutionary process.

Microwave sources have evolved continuously from large, hot devices with a lot of mechanical parts that are often difficult to fabricate into electronic and solid-state components. The motivations have been, almost exclusively, cost (broadly defined) and reliability. Of course, performance is also a factor; however, in many cases, it has been quite acceptable to sacrifice performance for cost and reliability. A perfect example is the move from TWT to solid-state transmitters on spacecraft. The power of early solid-state amplifiers could not match that of tubes, and the efficiency was often no better. But the need to improve reliability and reduce size (for greater functionality within weight limits) was so great that solid-state amps were simply made to work: improve the receivers and antennas a bit to pick up a couple dB, accept a small reduction in link margin, and it all works.

Indeed, most tubes and many of the older semiconductor devices had a cost or reliability floor that simply could not be pushed any lower. When there appears to be no technological route around such limitations, the pressure to develop—and to adopt—new technologies becomes irresistible. That's exactly what happened.

A continual problem in the early days of microwave solid-state sources was the lack of any concrete design procedure, to say nothing of circuit simulation capability. Producing a Gunn or IMPATT oscillator, or a varactor frequency multiplier, for that matter, was like trying to catch a very angry cat in a dark cellar. It involved many hours at a lab bench, tweaking the beast into existence. The designer, like the cat catcher, rarely emerged unscathed.

Indeed, the lack of circuit-level models for devices could be cited as a reason for the lack of systematic design methods. Such models could have helped. However, without computer circuit-simulation technology or other supporting technologies, such as accurately de-embedded microwave measurements, they would not have been terribly useful. I remember engineers

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in this era being openly contemptuous of computers, not entirely without reason. Using a computer in the 1960s and 70s was an exercise in slow motion that often could be outrun by some judicious lab work.

An important lesson of this experience is that the lack of a perfect, complete technology doesn't necessarily prevent its development. A curious aspect of technological advance is evident the way that a complete technology often arrives late in the game, frequently when we are ready to move on to new ones. This emphasis on the pragmatic and a willingness to work, often in the dark, with whatever is available, has resulted in extraordinarily fast technological progress. Compare this, say, to medical research, where the mindset is invariably first to understand the theory, then hope that practical results arise from it. We technogeeks are a lot faster, and, I dare say, more successful. ■

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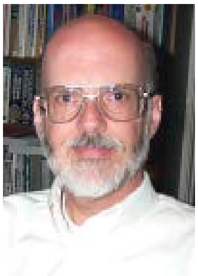
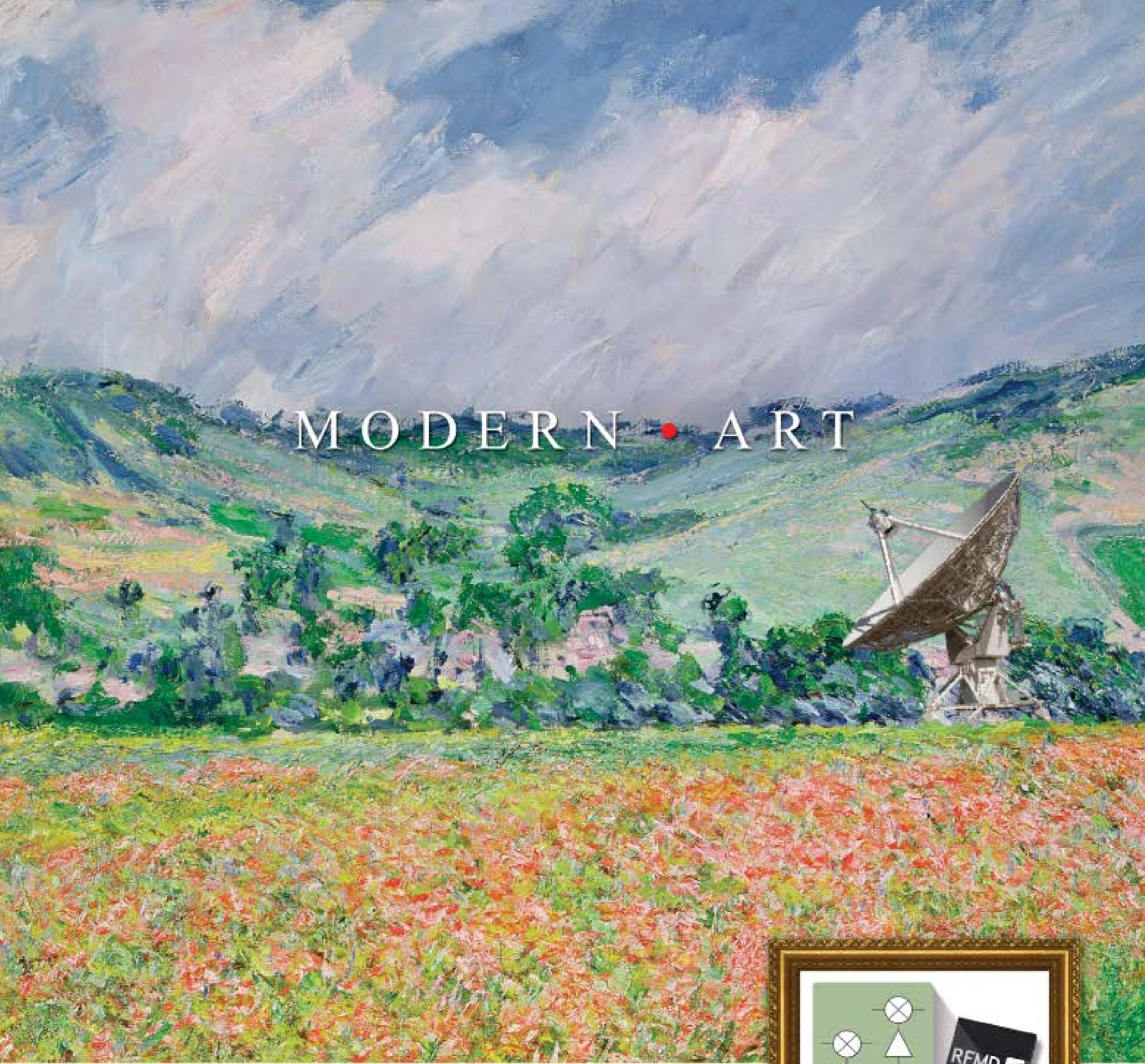
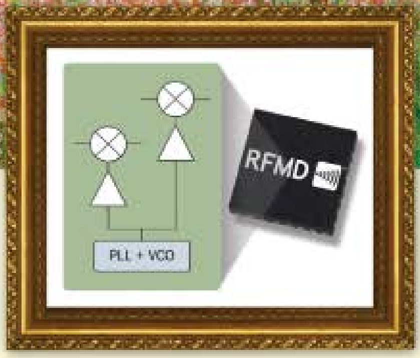


Stephen Maas *received his BSEE and MSEE degrees in electrical engineering from the University of Pennsylvania in 1971 and 1972, respectively, and his PhD degree in electrical engineering from UCLA in 1984. Since then, he has been involved in the research, design and development of low-noise and nonlinear microwave circuits and systems at the National Radio Astronomy Observatory (where he designed the receivers for the Very Large Array), Hughes Aircraft Co., TRW, the Aerospace Corp. and the UCLA Department of Electrical Engineering. Subsequently, he worked as an engineering consultant and founded Nonlinear Technologies Inc., a consulting company, in 1993. He is currently chief scientist of Applied Wave Research Inc. (AWR). He is the author of* Microwave Mixers *(Artech House, 1986 and 1992),* Nonlinear Microwave Circuits *(Artech House, 1988; second edition 2003),* The RF and Microwave Circuit Design Cookbook *(Artech House, 1998), and* Noise in Linear and Nonlinear Circuits *(Artech House, 2005). From 1990 until 1992 he was the editor of the* IEEE Transactions on Microwave Theory and Techniques *and from 1990 to 1993 was an Adcom member and publications chairman of the IEEE MTT Society. He received the Microwave Prize in 1989 for his work on distortion in diode mixers and the MTT Application Award in 2002. He is a Fellow of the IEEE.*





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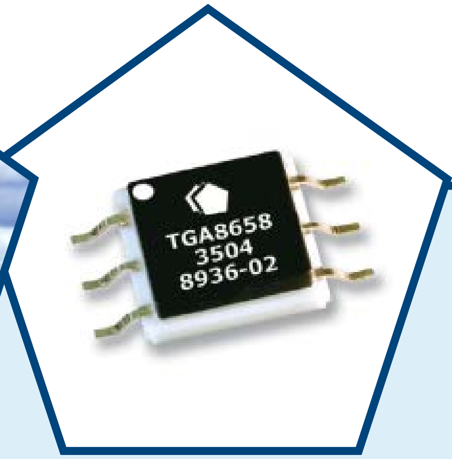

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HPA	7 - 8.5	(38) / –	21	– / 42	7 / 2000	Die	TGA2701
Driver Amp, SB	11 - 17	17 / –	23	6 / –	6 / 75	SM-06-12	TGA2507-SM
HPA	12 - 19	29 / –	25	–	5 - 7 / 435	SM-06-12	TGA2508-SM
2W HPA	12.5 - 16	(32) / 37	32	–	6 - 7 / 680	SM-01-24	TGA2503-SM
2W HPA	12.5 - 17	(33.5) / –	25	– / 25	7.5 / 650	SG-A1-6	TGA2510-EPU-SG
4W HPA	13 - 15	(36) / 41	25	–	7 / 1300	FL-A1-10	TGA8659-FL
6.5W HPA	13 - 16	(38) / –	24	–	8 / 2600	FL-A2-10	TGA2514-FL
2W HPA, PD	13 - 17	(34) / 38.5	26	– / 30	7.5 / 650	SG-A1-6	TGA2902-SCC-SG
2W HPA	13 - 17	(34) / 40	33	–	5 - 8 / 680	SG-A1-6	TGA8658-EPU-SG
K Band HPA	17 - 20	30 (32) / (42)	20	–	7 / 825	Die	TGA4530
Driver Amp	17 - 24	22 / –	19	4 / –	5 / 270	SM-09-16	*TGA2521-SM**
HPA, AGC, PD	17 - 24	(29) / 38	22	–	5 / 712	SM-010-20	*TGA2522-SM**
HPA	17 - 27	29 (31) / 37	22	–	7 / 760	Die	TGA4502-SCC
Gain Block & 2x/3x Multi	17 - 40	18 (22) / 24	22	7 / –	5 / 140	SM-A3-16	*TGA4031-SM**
HPA	25 - 31	35.5 (36) / –	22	–	6 / 2100	CP-A1-8	TGA4905-CP
MPA	25 - 35	25 / –	18	–	6 / 220	SM-A4-20	TGA4902-SM
7W HPA	26 - 31	(38.5) / –	22	–	6 / 4200	CP-A3-8	TGA4915-EPU-CP
2W HPA	27 - 31	32.5 (33) / –	22	–	6 / 840	CP-A2-8	TGA4513-CP
1W HPA	28 - 31	30 / –	19	– / 25	6 / 420	SM-A4-20	TGA4509-SM
4W HPA	28 - 31	36 (36.5) / –	22	– / 22	6 / 1600	Die	*TGA4906**
7W HPA	28 - 31	(38.5) / –	22	– / 20	6 / 3200	Die	*TGA4916**
Driver Amp	29 - 31	16 (17) / 22	15	–	6 / 1960	SM-A4-20	TGA4510-SM
MPA	33 - 47	27 (27.5) / 36	18	–	6 / 400	Die	TGA4522
HPA	36 - 40	30 / –	14	–	6 - 7 / 500	Die	TGA1171-SCC

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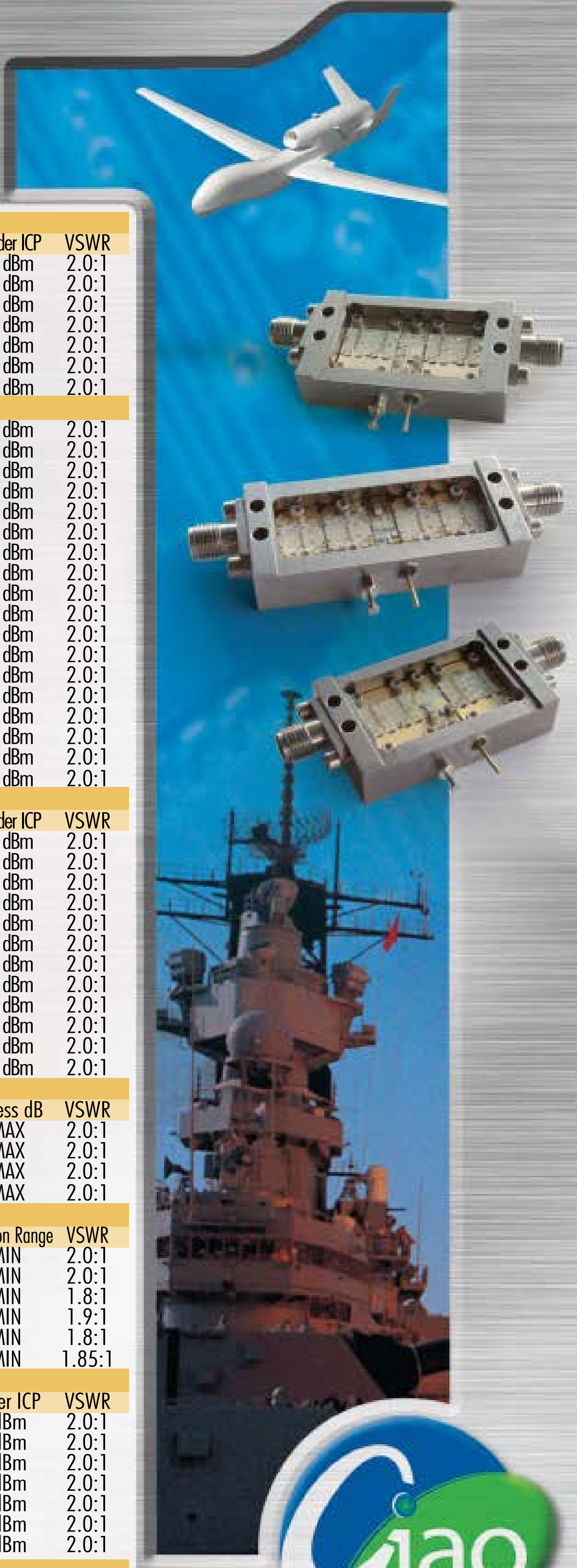

OCTAVE BAND LOW NOISE AMPLIFIERS

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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Harris and Cisco Team to Enhance Secure Wireless Networking

Harris Corp., an international communications and information technology company, and Cisco, the worldwide leader in networking, have teamed to speed the installation of Type 1 secure wireless networks for intelligence and civilian agencies of the federal government. The two companies combined their core strengths to provide Secure Wireless Architecture Type 1—a solution that expands the ability of federal agencies to view and share classified data over wireless information technology networks. “Harris and Cisco have industry-leading reputations for reliable, secure, mission-critical assured communications,” said Richard Rzepkowski, vice president of communications security products, Harris RF Communications. “Federal agencies can now obtain the best of Harris and Cisco products and expertise to develop premier Type 1 wireless solutions that meet their requirements for data confidentiality and integrity, user and device authentification, asset tracking and intrusion detection.” The combined Harris and Cisco solution includes Harris SecNet 54® device, the fastest Type 1 wireless encryption product available today, along with networking engineering, installation, management, support and training through the Harris Information Services business. Cisco will provide routers and controllers and other networking equipment along with networking design and engineering services. “In the past, the number of Type 1 wireless solutions was limited at the federal level due to policy mandates for wireless networking and other protocols governing access to classified data,” Rzepkowski said. “Harris and Cisco offer a complete solution that is backed by experts in wireless technology, system integration, installation and support, and the highest level of information security, along with a broad distribution network.”

Lockheed Martin Successfully Launches 6th Modernized GPS Satellite

A US Air Force modernized Global Positioning System Block IIR satellite, built by Lockheed Martin, was successfully launched from Cape Canaveral Air Force Station aboard a United Launch Alliance (ULA) Delta II launch vehicle. The satellite, designated GPS IIR-19M, is the sixth in a line of eight GPS IIR satellites that Lockheed Martin Navigation Systems, Valley Forge, PA, has modernized for its customer, the Global Positioning System Wing, Space and Missile Systems Center, Los Angeles Air Force Base, CA.

This mission represented the third successful launch of a GPS IIR-M satellite in under five months and is one of the final three Block IIR-M satellites planned for launch in 2008 to sustain and improve the GPS constellation. Each IIR-M satellite includes a modernized antenna panel that provides increased signal power to receivers on the ground, two new military signals for improved accuracy, enhanced encryption and anti-jamming capabilities for the military, and a second civil signal that will provide users with an open access signal on a different frequency.

“All of us at Lockheed Martin are proud of our longstanding partnership with the Air Force and the block IIR-M’s impressive record of performance,” said Don DeGryse, Lockheed Martin’s vice president of navigation systems. The GPS constellation provides critical situational awareness and precision weapon guidance for the military and supports a wide range of civil, scientific and commercial functions—from aircraft traffic control to the Internet—with precision location and timing information.

Lockheed Martin and its navigation payload provider ITT of Clifton, NJ, designed and built 21 IIR spacecraft and subsequently modernized eight of those spacecraft designated Block IIR-M for the Air Force. The final satellite, which includes a new demonstration for the new civil signal, known as L5, has just completed final integration testing and is on track for shipment to Cape Canaveral in preparation for launch in June.

The Global Positioning System enables properly equipped users to determine precise time and velocity and worldwide latitude, longitude and altitude to within a few meters. Air Force Space Command’s 2nd Space Operations Squadron (2SOPS), based at Schriever Air Force Base, CO, manages and operates the GPS constellation for both civil and military users.

Lockheed Martin is also leading a team which includes ITT and General Dynamics in the competition to build the US Air Force’s next-generation Global Positioning System, GPS Block III. The next-generation program will improve position, navigation and timing services for the warfighter and civil users worldwide and provide advanced anti-jam capabilities yielding improved system security, accuracy and reliability. A multi-billion dollar development contract is scheduled to be awarded by the Global Positioning System Wing, Space and Missile Systems Center, Los Angeles, CA, in early 2008.

Raytheon Missile and Radar Play Critical Roles in Satellite Intercept

Raytheon Co. played a pivotal role in the February 20th successful intercept of a non-functioning satellite. The company’s Standard Missile-3 was specially modified for its unique operation, performing beyond its intended capabilities to intercept the target 153 miles over the Pacific Ocean. Much engineering and technical expertise went into modifying the software on three SM-3 missiles for this one-time mission. Throughout the process, Raytheon engineers worked closely with its customer to ensure mission success.

At the same time, the Sea-based X-band radar, designed and built by Raytheon, tracked the satellite prior to the missile engagement and performed the hit assessment afterward. The radar performs the critical functions of cuing, tracking and discriminating a target. Its home port is Adak, Alaska, located approximately midway along the Aleutian Islands chain.

Northrop Grumman Wins US Army Missile Interceptor System Contract

Northrop Grumman Corp. is one of two companies awarded a contract by the US Army to design and demonstrate a prototype missile interceptor weapon system that will defend warfighters against rocket, artillery and mortar (RAM) threats. As part of the first phase of a multiphase program spanning five years, Northrop Grumman will design, fabricate, integrate and test hardware and software for a new battle element (BE) that is part of the Extended Area Protection and Survivability Integrated Demonstration (EAPS ID) program. The purpose of the EAPS ID is to create a mobile missile system (or BE) that can engage multiple, in-flight RAM threats accurately and protect forward deployed forces over a large defended area at a much lower system cost, and cost-per-hit, than is currently possible.

"The EAPS ID program is another opportunity that builds on Northrop Grumman's leadership in force protection systems and 50 years of missile systems excellence," said Frank Moore, vice president of the Missile Defense Division at Northrop Grumman's Mission Systems sector. "We are applying existing technologies with an innovative approach to demonstrate to the Army a prototype system that is low cost, mobile, accurate and able to defeat RAMs within this extended area of coverage, a much needed capability for our future force." The US Army Aviation and Missile Command/Research, Development and Engineering Center at Redstone Arsenal in Huntsville, AL, is managing this contract. Over the next five years, the Northrop Grumman team will demonstrate the technology for a complete kinetic energy weapon system battle element. The BE includes a low cost missile, a launcher, fire control radar and fire control computer to defeat a wide range of RAM threats. By the end of the contract, the Northrop Grumman team will have demonstrated this new capability to the Army during interceptor system testing at the Yuma Proving Ground, Yuma, AZ.

Northrop Grumman is leading an experienced technical missile and fire control team that includes Mitec Corp. and Torch Systems, both located in Huntsville. The initial award, option one, is for \$6 M and extends through November. Options two through five will be executed annually for a total contract value of up to \$40 M. ■


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ESA and EC Commit €624 M to Space Sentinels Project

The European Space Agency's director general Jean-Jacques Dordain and European Commission director general for Enterprise & Industry, Heinz Zourek, have signed an agreement, which establishes the allocation of an EC budget of €624 M to ESA as a contribution to the implementation of the GMES Space Component (GSC). This funding will be provided in two stages: €419 M for Segment 1 and €205 M for Segment 2 of the GSC programme.

The progressive implementation of GMES is made possible by the activities and investments of EU and ESA Member States. This is the second flagship initiative of the European Space Policy, following in the footsteps of the navigation system Galileo.

This agreement, together with the financial contributions from ESA Member States, will enable ESA to develop and launch the first three Sentinel satellites (Sentinel 1, 2 and 3), to set up the related ground segment for the reception, processing and dissemination to users of the satellite data (from the Sentinels and other satellites) and to undertake the development of further elements to come.

ESA's role is to implement the dedicated GMES Space Component, which involves developing the Sentinel satellite series and its ground segment, coordinating data access to the Sentinels and to other missions mainly from ESA Member States that contribute to fulfilling GMES service requirements.

MicReD Plays Major Role in NANOPACK Research

Flomerics' MicReD subsidiary will play a major role in the recently announced European-funded NANOPACK project, which aims to address thermal barriers to continued performance increases in semiconductors and power electronics by developing new technologies and materials for low thermal resistance interfaces and electrical interconnects.

The NANOPACK consortium, which is coordinated by Thales Research & Technology, will explore systems such as carbon nanotubes, nanoparticles and nano-structured surfaces using different contact-enhancing mechanisms combined with high-volume compatible manufacturing technologies such as electro-spinning. Recent groundbreaking work on nested channel interfaces will be utilized to exploit the beneficial properties of the new materials.

Within the NANOPACK project, MicReD and Budapest University of Technology & Economics (BUTE) will collaborate to create an intelligent micro-scale tester for thermal interface materials enabling thermal resistance measurements with at least one order of magnitude higher accuracy than is possible today. The project results will be disseminated through the organization of yearly public NANOPACK workshops in connection with the existing THERMINIC Workshops.

TASL and EADS Partner for Indian Army Bid

TATA Advanced Systems (TASL), a wholly owned subsidiary of Tata Industries, along with leading Tata technology companies (TATA Consultancy Services and TATA Power SED) is joining forces with EADS Defence & Security (EADS DS) to form a high tech team partnership to bid for the Indian Army's $1 B advanced tactical communications system project, which is expected to be announced later this year. Other technology partners include other TATA entities, Raytheon and Precision Electronics Ltd.

The proposed Indian Army Tactical Communications System is intended to replace the current AREN system and will make use of state-of-the-art technology. The new fully mobile communications system will be contemporary when fielded and will put the Indian Army on par with the most sophisticated tactical mobile systems currently in development for deployment around the globe. The partnership combines international lead systems integration expertise with local domain knowledge of the Indian market to provide an India Centric System with the latest technology and complete security.

Stefan Zoller, CEO of EADS Defence and Security, said, "India holds an important place on the world's stage and this programme recognises the need for India's armed forces to have the latest available technology. EADS Defence & Security and TATA can design and deliver one of the most sophisticated battlefield communications systems in the world, and at the same time, will make a significant contribution to India's high tech economy."

Anglo/German Partnership on LTE Base Stations

UK company picoChip is collaborating with German counterpart, mimoOn, to deliver the industry's first complete Long Term Evolution (LTE) base station reference design. The new PC86xx family of LTE reference designs cover the full range of eNode Bs from femtocells to multi-sector macrocells and is supported on the same common hardware platforms as picoChip's WiMAX products.

The LTE reference design empowers OEMs with faster time-to-market and lower development costs as they can focus their resources on adding value and differentiating their products. As a software-defined solution, it is future-proof and can be upgraded as the LTE standard is refined and updated. Finally, the common platform also supports WiMAX, with Wave 2 and MIMO available now, and upgrades for Release 1.x (FDD) and 16m.

"MimoOn's expertise enabled the development of this LTE base station solution—we are proud to be partnering with them," said Guillaume d'Eyssautier, president and CEO, picoChip. "It is crucial that we continue to provide our customers with the ability to develop systems on the latest wireless standards as they become available, to implement software upgrades as those standards are ratified, and to have the flexibility to improve them going forward. That is the picoChip chip advantage."

UK Scientists Investigate Cosmic Rays and Microchips

UK neutron scientists are tackling the challenge of cosmic radiation and its damaging effect on sensitive microchips in the aviation industry in the drive to develop more robust electronic equipment. Accelerated testing of microelectronic components at the Science and Technology Facilities Council's (STFC) ISIS neutron research centre replicates the effect of thousands of hours of flying time in just a few minutes.

Initial tests occurred at the end of 2006 at the ISIS neutron source in Oxfordshire, UK, and a £140 M new 'target station' or neutron source is in the final stages of completion alongside the original ISIS neutron source. The plan, subject to funding, is for the ISIS Second Target Station to include a dedicated and full-time instrument to test the effects of Single Event Effects (SEEs) and chip irradiation.

Results from this testing will allow manufacturers to mitigate against the problem and build triple redundancy into their electronic components. This increased confidence in the quality of electronic systems will help to make both civil and military aircraft safer.

Chris Frost, project leader of chip irradiation research at the ISIS neutron source, explained, "At ISIS we have the ability to produce intense beams of neutrons with similar energy ranges to those occurring naturally. This enables accelerated reliability testing of microelectric elements used in the aerospace industry. Once manufacturers understand where the biggest susceptibility problems lie, they can begin to redesign circuitry on a more robust basis." ■

Base Station Standardization: The New Innovation

One oft-cited maxim in the communications world states that standardization is the enemy of innovation: its premise is that once something is standardized there is, by definition, no room to alter or improve it and hence no remaining freedom in which to innovate. Indeed, this is frequently cited as an argument against standardization—the fact that kills innovation. What is perhaps more relevant is that once something is standardized, the market for that product can become commoditised; this is generally good for the customer or consumer and not so good for the profit margin of the supplier. This 'fact' is conveniently overlooked by those opposed to standardization in a particular area.

The Open Base Station Architecture Initiative (OBSAI) has produced a set of standards for the internal interfaces within a mobile communications base station. Far from killing innovation, the advent of the OBSAI standards and their widespread adoption in the industry has actually stimulated innovation and is enabling ground-breaking infrastructure solutions to find their way into OEM products much more quickly and easily than would otherwise have been the case.

In an OBSAI-based base station, such as the Flexi Base station platform from Nokia Siemens Networks, all of the interfaces are defined and these definitions, along with comprehensive test specifications, are publicly available to everyone to download from www.obsai.com, free of charge. This enables an innovative start-up company to design a set of requirements for which it knows there is a potential end-market and which it is able to comprehensively demonstrate compliance (by means of the test specifications). Such a start-up can therefore design a product, confident in the knowledge that it can quickly and easily be incorporated into any OEM base station that follows the OBSAI standard, either to offer a niche solution which the OEM may not have thought of, or to offer a significant benefit in a mainstream product area (e.g. improved PA efficiency). Since OBSAI is not restricted to a single OEM, this also helps to diversify the product development risk somewhat; a further important benefit to a start-up.

From an OEM perspective, it simplifies the testing of a potential new product from a third-party supplier, in a real base station scenario, as it eliminates the need to spend months educating said supplier in the details of a proprietary interface. In many cases, the OEM simply does not have the time or spare resources to undertake such an activity and this represents a significant barrier to entry for new technologies in the infrastructure marketplace.

The advent of the OBSAI specifications has allowed a number of start-up companies to develop innovative products for the base station space. One area, in particular, has benefited greatly from OBSAI's success: remote radio heads. Two start-ups that probably would not exist, but for the success of the base station standardization process are: Ubidyne and Axis Network Technology.

The Open Base Station Architecture Initiative is a forum of over 140 telecommunications companies—spanning module, component and base station vendor activity. Together they have created both open internal interface specifications for base station architecture and module specification covering the areas of Transport, Control, Baseband and Radio.

Ultra-wideband Equipment Shipments Will Exceed 400 Million

In 2006, it appeared that the market launch of ultra-wideband (UWB) was imminent. Several factors conspired to delay that, but ABI Research now expects UWB to see very strong growth starting in 2008, finding its first success in laptops and computer peripherals and eventually in mobile handsets. Forecasts indicate that shipments of UWB-enabled devices will grow from virtually nil today, to more than 400 million in 2013.

"The ultra-wideband market did not come out of the starting gate in 2006 as we had anticipated," says senior analyst Douglas McEuen. "There are several reasons for the delay, including a shakeout from three competing flavors of the technology to one, and the absence of global standards."

Now, however, conditions are ripe for a rapid takeoff. We are starting from "Year Zero": in 2007, only about 40,000 UWB-equipped devices shipped. This year, there will be perhaps a million, and ABI Research expects the curve to rise sharply thereafter. Because an official UWB standard has now been ratified in the US, North America is expected to lead this market for some time to come.

The current "sweet spot" in this market is UWB's application as a wireless USB enabler, connecting computers (especially notebooks) with printers, hard drives and other peripherals. An initial UWB "hub and dongle" configuration will enable users to retrofit the vast number of existing PCs and related equipment with wireless connections. UWB modules are just starting to appear in selected laptops (initially from Lenovo, Dell and Toshiba), but true silicon integration will take more time.

Later, other kinds of customer electronics such as digital cameras and camcorders, HDTVs and portable music devices will start to build the numbers, but, says McEuen, "Real market acceleration will only occur when UWB debuts in mobile handsets, where it will be used—possibly bundled with Bluetooth—to transfer music, pictures and video files. Even a small handset market penetration will deliver huge numbers. For UWB to see adoption in handsets, however, the price of the chipset must fall quite significantly."

"Ultra-wideband Connectivity" analyzes critical UWB market conditions, from drivers and obstacles to global regulations and standards. Key market semiconductor vendors are profiled. The study concludes with an in-depth market forecast that traces both positive and negative market forces numerically.

Mobile WiMAX More Opportunity Than Threat for Mobile Operators

From a mobile operator perspective, mobile WiMAX or 802.16e provides more of a service complement than a competitive threat, reports In-Stat. The mobile standard for WiMAX has been the subject of debate since its inception, the high-tech market research firm says. Mobile operators and vendors have disputed how this technology will impact their existing operations. The debate can be broken into two camps. "One camp, led by select equipment vendors with no stake in WiMAX has taken an either/or approach to discussing mobile WiMAX," says Daryl Schoolar, In-Stat analyst. "Any gain by WiMAX comes at the expense of other 3G data technologies. In the other camp, infrastructure vendors like Alcatel-Lucent, Motorola and Nokia Siemens see a world where multiple mobile wireless broadband technologies will co-exist. In-Stat believes that the latter camp's view will prevail."

Recent research by In-Stat found the following:

- IMT-2000 acceptance will open new markets for 802.16e.
- Mobile WiMAX will be more successful with laptop, external laptop adaptors and other consumer electronics than in phone handsets. It is these devices outside the handset where WiMAX provides the greatest competitive challenge to traditional cellular technologies.
- WiMAX will create new revenue opportunities for existing mobile operators.

The research, "Complement or Threat—WiMAX Strategies for Mobile Operators," covers the worldwide market for WiMAX and how 802.16e fits into the world of mobile operators. It discusses the impact of IMT-2000 acceptance of 802.16e as a 3G technology and examines three different mobile operators' strategies for WiMAX. It also provides device forecasts for both 802.16e and 3G. ■


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INDUSTRY NEWS

■ The **IEEE** International Reliability Physics Symposium's (IRPS) 46th annual conference will feature new technical sessions and invited papers on the latest microelectronic reliability issues. This year's IRPS takes place from April 27 through May 1, 2008, at the Hyatt Regency Phoenix at Civic Plaza in Phoenix, AZ. Keeping pace with the rapidly evolving challenges in the microelectronics industry, the IRPS introduces new sessions this year in nanoelectronic device reliability and thin film devices. Other key topics include assembly, interconnects, extreme environments, failure analysis, products and circuits, MEMS, memory, dielectrics (including transistor High-K), transistor, process integration, compound devices, soft-error rate (SER), electrostatic discharge (ESD) and latch-up. This year's technical program includes a record number of papers; more than 160 have been accepted for platform and poster presentations.

■ **RF Micro Devices** announced the completion of its acquisition of Filtronic Compound Semiconductors Ltd., a wholly owned subsidiary of Filtronic PLC. Under the terms of the transaction, RFMD paid an acquisition price of approximately £12.5 M in cash for Filtronic Compound Semiconductors. The acquisition price included the purchase of Filtronic Compound Semiconductors' six-inch GaAs wafer fabrication facility at Newton Aycliffe, United Kingdom, and the purchase of Filtronic Compound Semiconductors' millimeter-wave RF semiconductor business. RFMD expects the addition of Filtronic Compound Semiconductors' high-volume GaAs fab will significantly reduce RFMD's GaAs pHEMT manufacturing costs and provide incremental GaAs manufacturing capacity sufficient to support near-term anticipated growth. Additionally, RFMD expects the addition of Filtronic Compound Semiconductors' microwave and millimeter-wave component business will strengthen the product portfolio of RFMD's Multi-Market Products Group (MPG) and be accretive to RFMD's target margin profile for its multi-market business.

■ **Pendulum Instruments**, Stockholm, Sweden, announced the acquisition of **Rapco Electronics Ltd.**, a test and measurement company in Basingstoke, Hampshire, UK, specializing in high-precision time and frequency standards generation and distribution. Pendulum Instruments acquires all assets and operation of the privately owned Rapco company, by purchasing all outstanding shares. Rapco develops and produces precision time and frequency products, mainly for use by governmental, military, telecom, broadcasting and scientific customers. The product portfolio comprises high accuracy GPS-controlled master clocks, portable GPS-controlled frequency standards, time code generators, frequency and time-code distribution systems. Harald Kruger, Pendulum Instruments' CEO, says, "The inclusion of the Rapco product line with Pendulum Instrument's world-leading range of time and frequency products will strengthen our product portfolio and fill some gaps in the current program. The Rapco products fit perfectly in our worldwide distribution network."

■ **Planar Monolithics Industries** (PMI) announced its merger with **Planar Electronics Technology** (PET) and **Planar Filter Co.** (PFC). The synergy between these three Frederick, Maryland-based corporations enables the offering of upgraded hybrid MIC/MMIC RF and microwave components, subsystems and systems to their customers, as well as enhancing their ability to support major military and defense contractors.

■ **Microwave Communications Laboratories Inc.** (MCLI), St. Petersburg, FL, announced that it has successfully acquired its ISO9001:2000 Quality Certification. MCLI has been independently audited by SAI Global and assigned the Certification Number: QEC15974. This certification comes at a time when MCLI is celebrating it 20th anniversary as a manufacturer of precision RF components.

■ **Compel Electronics SpA**, Milan, Italy, a designer and manufacturer of interconnection systems and cable assemblies, celebrated the 10th anniversary of Compel Electronics GmbH, the company's German branch. Compel Electronics GmbH was set up in 1998 due to Compel Electronics SpA's need to develop and consolidate its presence in the interconnection market in Central and Eastern Europe. Over the years, Compel Electronics GmbH has established a series of important partnerships, such as with MP Systems, a Norwegian-based company, set up over 40 years ago that specializes in the development of Flexi-Racks and accessories, and Response Microwave, a Boston, MA-based company.

■ **AAE Systems Inc.** announced that it has expanded its operations with the opening of an office in India to provide regional sales, engineering and customer support. AAE says that it is confident that the geographic proximity of the new office to South Asia customers, seeking local engineering design and support services, will create even further business opportunities. The majority of the systems engineering design personnel will remain at AAE's headquarters in the US. Corporate functions, research and development, and manufacturing of AAE's Eclipse Digital Satellite Router product family will also remain at AAE's headquarters.

■ **Talley Communications Corp.** announced the opening of a new warehouse and sales facility in Dallas, TX. The new facility will expand Talley's logistic footprint to seven locations. This facility's primary focus will be to service contractors and carriers in the south central region of the United States. Andrew as well as CommScope cable and accessories will be stocked along with all necessary site hardware including tower steel, components and consumables.

■ **LPKF Laser & Electronics** has expanded its North American headquarters and moved to Tualatin, OR. The 18,000 sq. ft. space features administrative offices, a warehouse and product demonstration rooms. The expansion will help support the company's ongoing growth and help to further better service the company's North American customer base.

■ **Cadence Design Systems Inc.** announced that STMicroelectronics has successfully taped out its newest leading-edge mobile multimedia Nomadik platform using the Cadence Incisive Enterprise verification family. The advanced verification solution for verification engineers included the Incisive Specman Elite Testbench and Incisive Enterprise Manager, a powerful combination of technology within the Plan-to-Closure Methodology for automating the verification process. The STMicroelectronics Nomadik family of multimedia application processor chips enables portable devices to play music, take pictures, record video and host two-way video communication in real time. Many of ST's latest networking designs include ARM-based cores and use the internal STBus for much of the underlying infrastructure.

■ **Agilent Technologies Inc.** announced a comprehensive solution for monitoring and troubleshooting new mobile soft switch (MSS) networks and the Voice over IP (VoIP) services they deliver. Agilent's enhanced QoS analyzer and session trace applications further extend the assureME application portfolio by simplifying the process of managing MSS networks. Network and service quality is ensured through rapid identification and analysis of potential issues, and in-depth troubleshooting for early problem resolution. Deploying mobile soft switches is a critical step to ensure that mobile operators remain competitive in their key service, voice telephony. The solution enables mobile operators to track a new extensive set of key performance indicators (KPIs), including voice quality, to manage soft switch infrastructure, and to deliver an acceptable service experience.

CONTRACTS

■ **Mercury Computer Systems Inc.** announced that its wholly owned subsidiary, Mercury Federal Systems Inc., has received a $2.5 M contract award from the US Army Ft. Monmouth's CECOM in New Jersey, for development and demonstration of a testbed to support the Joint Counter Radio-Controlled Improvised Explosive Device Electronic Warfare (JCREW) program. The testbed will be used by the US Government to develop and test advanced open-architecture technology for counter-IED systems development. Mercury also announced a related agreement with ITT Electronic Systems to demonstrate algorithm component portability for the JCREW testbed architecture, which is an important element of the R&D objectives of the program.

■ ***n*Gimat** has received a Small Business Innovation Research (SBIR) Phase II award from the US Navy to prototype a low cost, miniaturized tunable filter for use in X-band active electronically steered antenna (AESA) array-based radar systems. This is a two-year contract with a funding of $750,000. "We have developed an X-band tunable filter with a steep roll-off (> 30 dB at ±10% f_o), as well as moderate tunability and loss," said Zhiyong Zhao, program manager for *n*Gimat. "We will further perfect the design and improve the materials and processing, and a fully functional device will be delivered to the Navy by the end of Phase II."

■ **TRAK Microwave**, a Smiths Interconnect business, and University of South Florida (USF), College of Electrical Engineering, have jointly won a full matching grant from the Florida High Technology Corridor program. The grant sponsors research programs aimed at "Measuring and Modeling of Phase Noise Conversion in Amplifiers and Frequency Multipliers." This is the second phase of the project started several years ago to advance the test, measurement and modeling relationships between 1/f noise and phase noise conversion in active microwave and millimeter-wave circuits.

■ **Tyco Electronics** announced it has received a $2 M contract award from Raytheon Missile Systems to produce products for Raytheon's precision-guided, long-range Excalibur artillery projectile. Tyco Electronics will produce M/A-COM telemetry transmitter modules, Global Positioning System (GPS) antennas and telemetry antennas to support Excalibur and contribute to the projectile's accuracy in both urban and complex terrain and reduce collateral damage.

■ Eutelsat Communications has awarded **Thales Alenia Space** the contract for the W3B communications satellite. As prime contractor, the company will be in charge of the design, manufacturing, test and delivery of the satellite. Planned for a launch during the second quarter of 2010, the satellite will provide broadcast and telecommunications services over Europe, the Middle East and Africa. Designed with a lifetime of more than 15 years, W3B will provide three Ka-band and 53 Ku-band transponders. It will be equipped with three deployable and two fixed antennas, have a maximum launch mass of 5.4 tons and will deliver more than 12 kW of electrical power at the end of life.

■ Following Thailand's purchase of an integrated air surveillance system from Sweden in a government to government deal, FMV, the Swedish Defense Material Administration, has placed a contract with **Saab** valued at $310 M for the development and production of the systems. Deliveries will take place during 2011. The order comprises six Gripen fighters of the C/D version, two Saab 340 aircraft, of which one will be equipped with the company's radar surveillance system Erieye, and associated equipment and services. Also included in the agreement is a command and control system, which will be the link between the Erieye system and the Gripen fighters. The complete system will be used for air surveillance and protection of Thailand's territory.

PERSONNEL

■ **Tony Belotis**, co-founder of GT Microwave, passed away on February 20, 2008. A proud US Navy veteran,

and a member of the Association of Old Crows, Belotis had worked at CTI, Triangle, Merrimac, AT&T and Flexco prior to starting GT Microwave.

▲ *Roberto Pinto da Silva*

■ **Roberto Pinto da Silva**, formerly vice president of sales for Radio Frequency Systems (RFS) Latin America since 2005, has replaced Luis Antonio Alves de Oliveira as president of RFS Brazil. Oliveira leaves RFS to pursue new endeavors outside the telecom industry. RFS Brazil's new president counts more than 30 years of experience in the technology business. Over 26 years, Pinto rose through the ranks of IBM and gained responsibility for the company's entire line of medium-sized computers in Brazil. He oversaw the development of many business partnerships, including joint ventures, as well as technology and production licenses. In 1998, he was hired by Allen Telecommunications of Dallas, TX, to coordinate and implement the launch of the company's production and sales operations in all of South America. Pinto's international expertise is the result of his five-year tenure in the United States (where he acted as a technical liaison, business negotiator and training expert), coupled with his extensive travels in Europe, Asia and South America.

▲ *Ehud Netzer*

■ Orbit's board of directors has appointed **Ehud Netzer** as president and CEO of Orbit Technology Group. Prior to joining Orbit, Netzer held the position of vice president and head of Rafael Advanced Defense Systems division, the second largest division with over 1000 employees. During this period, the sales of the division increased by 20% annually. Previously, Netzer held senior positions at Rafael, including: head of naval systems, intelligence and EW, head of the naval combat systems line, and head of the dynamics and control section. In addition, he served as director for Opgal, a subsidiary of Rafael and Elbit Systems. Netzer holds a PhD degree in mechanical engineering from Stanford University and BA and MSc degrees in mechanical engineering from Technion University in Israel. Netzer has also participated in the advanced management program at Harvard Business School. He replaces Shlomo Yariv, who was Orbit's president and CEO since 2003.

■ Strategy Analytics announced the promotion of **Chris Ambrosio** to executive director, wireless practice. Ambrosio's experience is focused on device manufacturer issues in the areas of strategic business planning, market assessment and competitive analysis. In addition to his experience working with top global handset suppliers, he has also worked intensively with service providers, semiconductor vendors and software suppliers in areas of cellular product planning and enabling technologies assessments.

Broadband Amplifiers by AML Communications

To Order Call: (805)388-1345 ext. 203

Model	Frequency (GHz)	Gain (dB)	Flatness (dB) max	NF (dB) max	P1dB (dBm) min	VSWR (In/Out)	DC Current @ +12/+15VDC
Broadband Low Noise Amplifiers							
AML016L2802	0.1 – 6.0	28	±1.25	1.3*	+7	2.0:1	190
AML48L3001	4.0 – 8.0	30	±1.0	1.2	+10	1.8:1	150
AML412L3002	4.0 – 12.0	30	±1.5	1.5	+10	1.8:1	150
AML218L0901	2.0 – 18.0	9	±1.0	2.2	+5	2.5:1	60
AML0518L1601-LN	0.5 – 18.0	16	±1.0	2.7	+8	2.2:1	100
AML0126L2202	0.1 – 26.5	22	±2.25	3.5*	+8	2.2:1	170
AML1226L3301	12.0 – 26.5	33	±2.0	2.8	+8	2.5:1	200
Broadband Medium Power Amplifiers							
AML0016P2001	0.01 – 6.0	21	±1.25	3.2*	+23*	2.0:1	480
AML26P3001-2W	2.0 – 6.0	28	±2.5	6	+33	1.8:1	1500
AML28P3002-2W	2.0 – 8.0	30	±2.0	5.5	+33	2.0:1	2000
AML218P3203	2.0 – 18.0	32	±2.5	4	+25	2.0:1	450
AML618P3502-2W	6.0 – 18.0	35	±2.5	4	+33	2.0:1	1850
Narrow Band Low Noise Amplifiers							
AML23L2801	2.8 – 3.1	28	±0.75	0.7	+10	1.8:1	150
AML1414L2401	14.0 – 14.5	24	±0.75	1.5	+10	1.5:1	130
AML1718L2401	17.0 – 18.0	24	±0.75	1.6	+10	1.8:1	150

Low Phase Noise Amplifiers — **Phase noise (dBc/Hz) at offset**

Part Number	Frequency (GHz)	Gain (dB)	Output Power (dBm)	100Hz	1KHz	10KHz	100KHz
AML811PN0908	8.5 – 11.0	9	17	-154	-159	-167	-170
AML811PN1808	8.5 – 11.0	18	18	-152.5	-157.5	-165.5	-168
AML811PN1508	8.5 – 11.0	15	28	-145.5	-153.5	-158.5	-164.5
AML26PN0904	2.0 – 6.0	9	20	-150	-165	-165	-178
AML26PN1201	2.0 – 6.0	11	15	-155	-160	-160	-175

High Dynamic Range Amplifiers

Part Number	Frequency (MHz)	Gain (dB)	P1dB (dBm)	OIP3 (dBm)	DC
AR01003251X	2 – 32	21	32	52	+28V @ 470mA
AFL30040125	50 – 500	23	28	53	+28V @ 700mA
BP60070024X	20 – 2000	32	30	43	+15V @ 1100mA

*Above 500MHz.

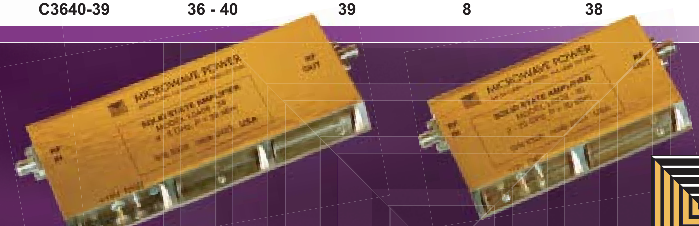
Power Amplifiers by Microwave Power

To Order Call: (408)727-6666 ext. 42

Model	Frequency (GHz)	Psat (dBm)	Psat (W)	P1dB (dBm)	Gain (dB)	DC Current(A) @ +12V or +15V
Broadband Microwave Power Amplifiers						
L0104-43	1 - 4	42.5	17.8	41.5	45	14
L0204-44	2 - 4	44	25	42.5	45	14
L0206-40	2 - 6	40	10	38.5	40	8.5
L0208-41	2 - 8	41	12	40	40	17
L0218-32	2 - 18	32	1.4	31	35	5
L0408-43	4 - 8	43	20	41.5	45	17
L0618-43	6 - 18	43	20	41.5	45	22
L0812-46	8 - 12	46	40	45	45	28
Millimeter-Wave Power Amplifiers						
L1826-34	18 - 26	34	2.5	33	35	4
L1840-27	18 - 40	27	0.5	26	30	2
L2240-28	22 - 40	28.5	0.7	27	30	3
L2630-39	26 - 30	39	8.0	38	40	15
L2632-37	26 - 32	37	5.0	36	38	10
L2640-31	26 - 40	31	1.2	30	30	5
L3040-33	30 - 40	33	2.0	32	33	9
L3337-36	33 - 37	36	4.0	35	40	12
L3640-36	36 - 40	36	4.0	35	40	10

High-Power Rack Mount Amplifiers

Model	Frequency (GHz)	Psat (dBm)	Psat (W)	P1dB (dBm)	Pac (kW)	Height (in)
C071077-52	7.1 - 7.7	52.5	170	51.5	1.8	10.25
C090105-50	9 - 10.5	50	100	49	1	8.75
C140145-50	14 - 14.5	50.5	110	49.5	2	10.25
C1416-46	14 - 16	46	40	45	0.35	5.25
C1820-43	18 - 20	43	20	41.5	0.25	5.25
C2326-40	23 - 26	40	10	39	0.25	5.25
C2630-45	26 - 30	45	30	44	0.45	5.25
C3236-40	32 - 36	40	10	39	0.25	5.25
C3640-39	36 - 40	39	8	38	0.24	5.25



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■ Rohde & Schwarz announced that **Jack Cowper** has been appointed president and COO of Rohde & Schwarz Inc., reporting to CEO Wolfgang Schmittseifer. He will also retain the position of president of Rohde & Schwarz, Canada, which he has held since June 2007. In his new position, he will be responsible for all operational areas of Rohde & Schwarz in North America. Cowper has more than 15 years of experience in the test and measurement industry. He joined Rohde & Schwarz in 2000 and has served as marketing manager, director of marketing and as vice president during his tenure with the company. Before joining Rohde & Schwarz, Cowper served in product and strategic marketing, and corporate management positions at Tektronix and Applied Microsystems. He received his BSEE degree from Pennsylvania State University, his MBA degree from the University of Oregon, and has completed the Harvard University Program for Management Development.

■ Aperto Networks, a builder of WiMAX base stations and subscriber units, announced the appointment of **Ruchir Godura** as vice president, global customer service and South Asia sales. Godura brings over 14 years of experience in the telecom industry to Aperto. Over his career in India and North America, he has managed sales, customer service, and product management for a variety of communication technologies including WiMAX, CDMA, DSL, Metro Ethernet and IPTV. Most recently, Godura was CMO, Enterprise Services, at Bharti Airtel, India's largest GSM operator. Before Airtel, Godura helped establish the WiMAX business at Telsima as head of worldwide sales. Prior to that, he spent nearly nine years with UTStarcom, setting up the company's global customer support services in New Jersey before moving to India to lead its South Asia operations. Godura holds a BS degree in computer science and engineering from the Indian Institute of Technology, Delhi, and a MS degree in computer science from the University of Delaware.

■ Technical Communities, a service provider for technical organizations that sell to US government agencies, military organizations and prime federal contractors, announced that **Jeremiah Cunningham** is joining the company as a vice president, business development. Cunningham, who will be based in Technical Communities' Washington, DC regional office in Herndon, VA, comes to Technical Communities from CDW-G, where he was the director of strategic sales. As a vice president, business development, Cunningham will assume responsibility for a strategic approach to the company's continued business expansion of GSA distribution of IT products and services. He will be directly responsible for the establishment of agreements between Technical Communities and manufacturers as well as ongoing sales strategy.

■ Peregrine Semiconductor Corp., a supplier of RF-CMOS and mixed-signal communications ICs, announced the appointment of former Qualcomm executive **Jeffrey Belk** to its board of directors. Belk brings a wealth of industry experience to Peregrine's Board from his 14-year tenure at Qualcomm, where he held a variety of senior

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SBTC-2-25+	1000-2500	50 Ω	3.49
SBTC-2-10-75+	10-1000	75 Ω	3.49
SBTC-2-15-75+	500-1500	75 Ω	3.49
SBTC-2-10-5075+	50-1000	50/75 Ω	3.49
SBTC-2-10-7550+	5-1000	50/75 Ω	3.49
SCA-4-10+	5-1000	50 Ω	6.95
SCA-4-10-75+	10-1000	75 Ω	6.95
SCA-4-15-75+	10-1500	75 Ω	7.95
• SCA-4-20+	1000-2000	50 Ω	7.95

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management positions, including his most recent role as senior vice president of strategy and market development. In this position, Belk examined changes in the wireless ecosystem and formulated approaches to help accelerate mobile broadband adoption and growth.

REP APPOINTMENTS

■ **Aethercomm Inc.** appointed Rick Biggs, **BA-Sales**, Denton, TX, as its exclusive sales representative covering Texas, Oklahoma, Arkansas and Louisiana. BA-Sales will represent and support Aethercomm's entire product line offering. BA-Sales can be reached at (972) 679-5871, www.ba-sales.com.

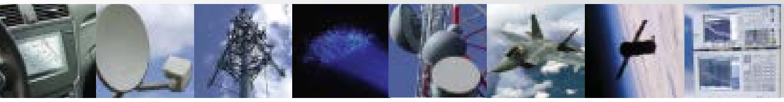
■ **Allied Electronics**, a subsidiary of Electrocomponents plc, has signed an agreement to distribute **Quest Technology**'s new LightStar fiber optic products. Allied is currently the exclusive catalog distributor of Quest's entire LightStar product line, which features new epoxyless field-terminated fiber optic connectors featuring patented impact mount technology.

■ **Sprague-Goodman Electronics Inc.** has appointed **JLF Engineering Inc.**, Spencer, MA, to represent its line of trimmer capacitors, transformers, fixed and variable inductors and tuning tools in Connecticut, Maine, Massachusetts, New Hampshire and Vermont. JLF represents a wide selection of RF/microwave products and companies throughout the Northeast. Contact JLF Engineering at (508) 769-5035, jlfengineering@charter.net.

■ **MU-DEL Electronics Inc.** has signed two new representatives to market its RF multicouplers, RF frequency converters, RF switching distribution systems and specialized RF products. **Kelley Systems Co.** (KSC), Lake Guntersville, AL, will be supporting MU-DEL in Alabama, Tennessee and Mississippi. Contact KSC at (256) 582-8677. **Tech Marketing Associates Inc.** (TMC), Mountain View, CA, will be supporting MU-DEL in Northern California and northern Nevada. Contact TMA at (408) 736-3687 or (650) 968-0102.

■ The **Evans Capacitor Co.** has engaged René Terrien of **Firadec**, St.-Nazaire, France, to represent its products in Europe. Representing a full line of capacitors, Terrien brings a breadth of experience to capacitor design and selection. Contact him at +33 2 40 01 26 51 or rene.terrien@firadec.fr.

MMICs FOR M



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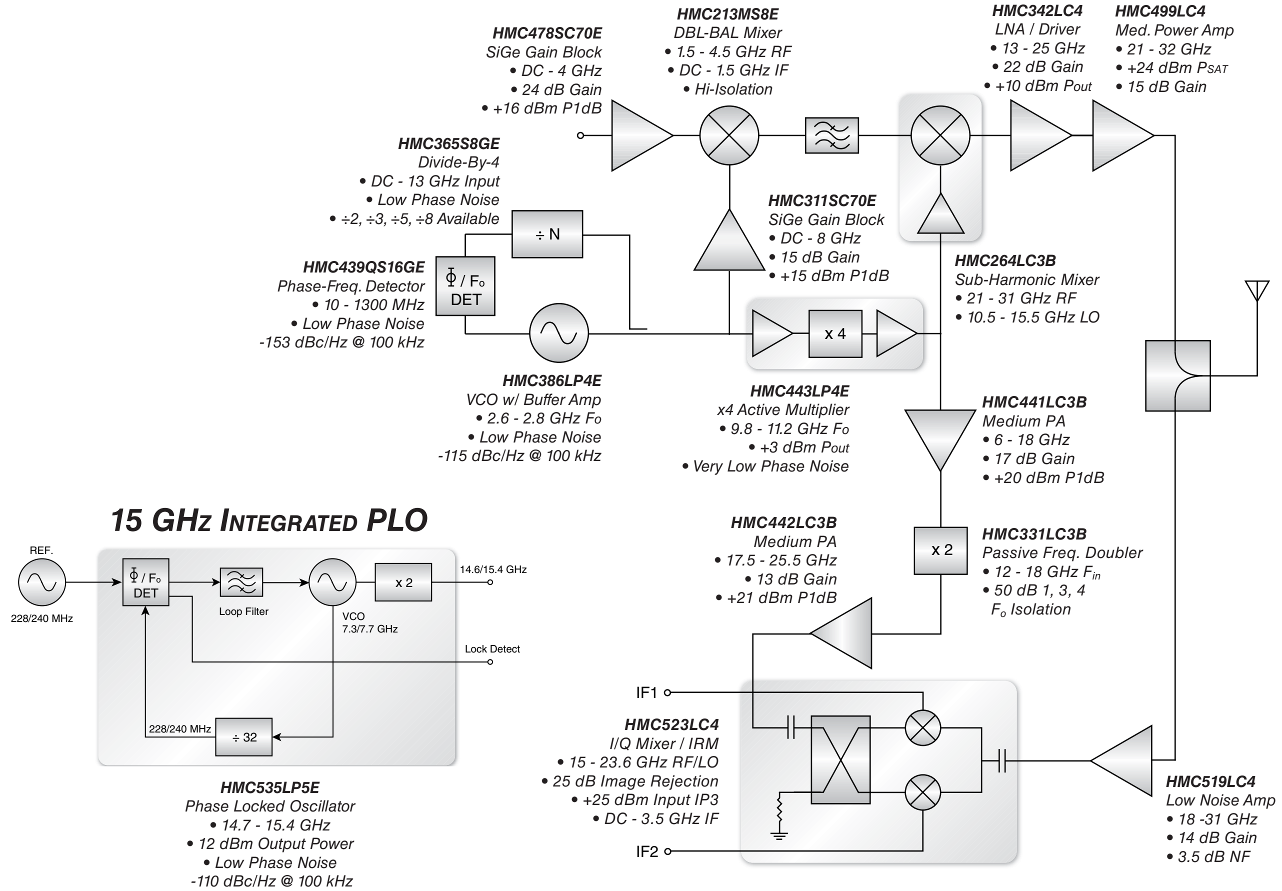

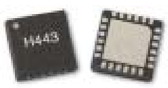

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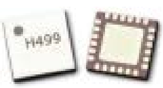

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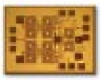
Driver Amplifier	HMC441LP3E HMC451LC3 HMC516LC5	HMC441LP3E HMC451LC3 HMC516LC5	HMC441LC3B HMC451LC3 HMC490LP5E	HMC383LC4 HMC442LC3B HMC498LC4	HMC383LC4 HMC442LC3B HMC498LC4	HMC383LC4 HMC283LM1 HMC499LC4	HMC383LC4 HMC-APH403 HMC-APH473 HMC-APH510	HMC-ABH209 HMC-ABH241 HMC-ABH403	HMC-AUH317 HMC-AUH318 HMC-AUH320
Power Amplifier	HMC486LP5E HMC590LP5E HMC591 HMC591LP5E	HMC487LP5E HMC592 HMC608 HMC608LC4	HMC489LP5E HMC592 HMC-APH462	HMC-APH196 HMC-APH462 HMC-APH478 HMC-APH596	HMC498LC4 HMC-APH462 HMC-APH518 HMC-APH596	HMC499LC4 HMC-APH196 HMC-APH460 HMC-APH608	HMC283LM1 HMC-APH473 HMC-APH510 HMC-AUH256	HMC-ABH209 HMC-ABH241 HMC-ABH403	HMC-AUH317 HMC-AUH318 HMC-AUH320
Wideband (Distributed) Amplifiers	HMC634LC4 HMC-AUH232 HMC-AUH249 HMC-AUH312	HMC633LC4 HMC-ALH435 HMC-ALH444 HMC-ALH482	HMC606LC5 HMC-ALH102 HMC-ALH476 HMC-AUH312	HMC463LH250 HMC-ALH216 HMC-ALH435 HMC-ALH445	HMC635 HMC-ALH311 HMC-ALH476 HMC-AUH256	HMC-ALH140 HMC-ALH244 HMC-ALH313 HMC-ALH364	HMC-ALH310 HMC-ALH313 HMC-ALH369 HMC-ALH376	HMC-ALH376 HMC-AUH312	
Attenuator: Analog	HMC346LP3E	HMC346LP3E	HMC346LP3E	HMC-VVD102	HMC-VVD102	HMC-VVD102	HMC-VVD106	HMC-VVD106	HMC-VVD104
Divide-by-2	HMC361S8GE	HMC364S8GE	HMC492LP3E	HMC492LP3E					
Divide-by-4	HMC362S8GE	HMC365S8GE	HMC493LP3E	HMC447LC3	HMC447LC3	HMC447LC3			
Divide-by-8	HMC363S8GE	HMC363S8GE	HMC494LP3E						
Multiplier: Active X2	HMC368LP4E HMC575LP4E HMC561LP3E	HMC573LC3B HMC561LP3E HMC-XDB112	HMC573LC3B HMC561LP3E HMC-XDB112	HMC448LC3B HMC576 HMC576LC3B	HMC448LC3B HMC576 HMC576LC3B	HMC448LC3B HMC577LC4B HMC578LC3B	HMC449LC3B HMC578LC3B HMC579		
Multiplier: Active X4		HMC443LP4E	HMC370LP4E						
Multiplier: Passive X2	HMC189MS8E	HMC189MS8E	HMC204MS8GE	HMC204MS8GE	HMC205	HMC331	HMC331		
I/Q Receiver	HMC567LC5	HMC568LC5	HMC569LC5	HMC571 HMC571LC5	HMC572 HMC572LC5	HMC572 HMC572LC5			
I/Q Mixer / IRM	HMC520LC4 HMC525LC4 HMC620 HMC620LC4	HMC521LC4 HMC527LC4	HMC521LC4 HMC522LC4 HMC527LC4 HMC528LC4	HMC523 HMC523LC4	HMC523 HMC523LC4 HMC524 HMC-MDB172	HMC524 HMC524LC3B HMC-MDB172	HMC404 HMC555 HMC556 HMC-MDB172	HMC-MDB171 HMC-MDB207	
Mixer: Fundamental	HMC219MS8E HMC220MS8E HMC553LC3B HMC558LC3B	HMC411MS8GE HMC412MS8GE HMC553LC3B HMC558LC3B	HMC411MS8GE HMC412MS8GE HMC553LC3B HMC558LC3B	HMC260LC3B HMC292LC3B HMC554 HMC554LC3B	HMC260LC3B HMC292LC3B	HMC292LC3B HMC329LC3B HMC557LC4 HMC560LM3	HMC294 HMC329LM3 HMC560 HMC560LM3	HMC-MDB169	
Mixer: Sub-Harmonic				HMC258LC3B HMC258LM3	HMC264LC3B HMC338LC3B	HMC265LM3 HMC338LC3B	HMC338 HMC339	HMC-MDB218	
Switch	HMC547LP3E HMC641	HMC607 HMC641	HMC547LP3E HMC607	HMC547LP3E HMC641			HMC-SDD112	HMC-SDD112	
VCO & PLO: **Requires X2 or X4	HMC506LP4E HMC532LP4E HMC586LC4B HMC587LC4B	HMC515LP5E HMC534LP4E HMC582LP5E HMC588LC4B	HMC529LP4E HMC535LP5E HMC584LP5E HMC632LP5E	HMC429LP4E**	HMC431LP4E**	HMC515LP5E** HMC531LP5E**	HMC505LP4E HMC506LP4E		



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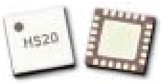

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PHASE NOISE: THEORY VERSUS PRACTICALITY

Modern electronics rely on fast low noise time bases for a variety of applications. As a result, oscillator designers must always push the limit of low noise design. One of the major hurdles in this endeavor is phase noise. Phase noise is an undesirable entity that causes distortion and high bit error rates. It is therefore important to understand this phenomenon and reduce its effects on higher level systems. This article discusses different time bases and their properties, crystal oscillator properties, phase noise definition and calculations, ways to reduce phase noise in oscillator design, real oscillator phase noise plots, and outside noise/interference effects on phase noise. It will be shown that phase noise is unavoidable but manageable. Through careful oscillator design, phase noise can be reduced, and through informed consumer system design can accommodate the real world oscillator.

Modern electronics use time bases for a wide variety of applications. Communications systems rely on time bases for modulation and demodulation of data, GPS systems rely on them for accurate positioning, and a host of other applications rely on time bases to manage the flow of data within their system. As the range of applications grows and the frequency increases, designers need time bases that are tighter in stability and lower in noise. As a result, oscillator designers need to continually push the limit of tight stability, low noise design.

One of the major issues facing oscillator designers is the phase noise phenomenon. Phase noise is an undesirable entity that is present in all real world oscillators and signal generators. It can cause distortion or complete loss of incoming information in traditional receivers and high bit error rates in phase modulated applications. This makes it a necessity to quantify and understand this noise in the oscillator, so that its effect on the higher-level product is minimized.

Time is a concept all humans innately grasp as a function of life. It defines our world into days, months and years. When we look at a clock we know what time it is. However, do we really understand time and how it is defined? Like all measurements, time is measured with a degree of uncertainty. In engineering applications we use time bases that are by nature imperfect, adding uncertainties into the applications that are driven by them. A byproduct of this uncertainty is phase noise. The inherent instability present in every oscillator manifests itself as a spectrum of noise around the frequency of the oscillator. This noise band is generally measured from the carrier to 1 MHz away from the carrier. It is depicted as a graph of dBc/Hz vs. f(Hz), which shows how far down from the carrier power the noise power is at a given frequency in the

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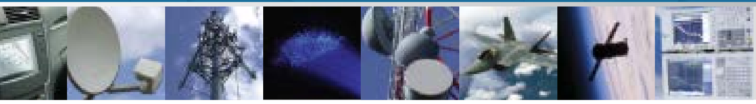

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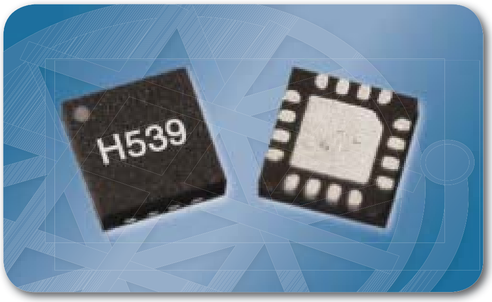

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	0.7 - 4.0	2-Bit Digital	0.9	4 to 12	54	0 / +3V	**HMC291E**
	DC - 6	3-Bit Digital	0.7	1 to 7	50	TTL/CMOS	**HMC468LP3E**
	DC - 5.5	4-Bit Digital	0.8	1 to 15	50	TTL/CMOS	**HMC540LP3E**
NEW!	DC - 6	4-Bit Digital, Serial & Parallel Control	2.5	45	50	0 / +5V	**HMC629LP4E**
	0.7 - 3.7	5-Bit Digital, Serial Control	2.1	1 to 31	48	Serial TTL/CMOS	**HMC271LP4E**
	0.7 - 3.8	5-Bit Digital	2.1	1 to 31	48	0 / +3V	**HMC273MS10GE**
	0.7 - 3.8	5-Bit Digital, Serial Control	1.5	0.5 to 15.5	52	Serial TTL/CMOS	**HMC305LP4E**
	0.7 - 3.8	5-Bit Digital	1.5	0.5 to 15.5	52	0 / +3V	**HMC306MS10E**
	DC - 3	5-Bit Digital	1.3	1 to 31	45	TTL/CMOS	**HMC470LP3E**
	DC - 4	5-Bit Digital	0.7	0.25 to 7.75	50	TTL/CMOS	**HMC539LP3E**
	DC - 3	6-Bit Digital	1.5	0.5 to 31.5	45	TTL/CMOS	**HMC472LP4E**
	DC - 3	6-Bit Digital, Serial Control	1.2	0.5 to 31.5	45	Serial TTL/CMOS	**HMC542LP4E**
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Distributed in the Americas by Future Electronics: 800 Future-1, ext. 2754 www.futureelectronics.com/rf following article. Many topics will be discussed, including:

- Different time bases and their properties
- Crystal oscillator properties
- Phase noise definition and calculations
- Ways to reduce phase noise in oscillator design
- Real oscillator phase noise plots
- Outside noise/interference effects on phase noise

The following information will show that phase noise is a manageable problem. Oscillator designers can work to minimize the phase noise in oscillators, and system designers using the oscillators can better design their systems and choose the correct oscillator for their needs. It is also clear that outside noise and interference degrades the phase noise performance of oscillators. Engineers need to be aware of these factors, so that their effects can be anticipated or avoided.

TIME BASES

Engineers are used to looking at a frequency counter and reading the display to determine at what frequency an oscillator is operating at. But what is the accuracy of that measurement? The uncertainty of the measurement is dependant upon the accuracy of the time base the counter uses to drive its internal circuitry. This time base is not perfect and therefore skews the results of the frequency reading. For those applications requiring very precise frequency measurements, the accuracy of this time base is of the utmost importance.

Time, like every other measurement man has ever made, has an uncertainty. However, time is the most accurate standard that mankind has produced. The second is defined as the resonant frequency of cesium 133 (Cs-133), which is 9,192,631,770 Hz.[8] Therefore, by measuring these vibrations of cesium, a standard for time is established. NIST (National Institute for Standards and Technology) is the keeper of the time standard for the United States. As of 2005, the NIST standard has an uncertainty of $5X10^{-16}$ s, which means it would neither gain nor lose a second in 60 million years.[9] Cesium clocks, despite their accuracy, have several drawbacks making them unsuitable for mass usage in commercial electronics. The cost of cesium standards is prohibitive for using them as a time base. Secondly, they are large. The NIST standard fills a good portion of a room, and while smaller cesium standards are readily available they are not a portable item. The cesium standard due to its nature has a warm-up time; therefore, keeping power supplied to it is important, because a break in power could mean degradation in accuracy. Lastly, cesium standards use fuel; the cesium naturally depletes, rendering the device effectively 'out of gas.'

Since cesium cannot effectively be used for commercial electronics or for most lab applications other timing sources must be considered. Rubidium offers extremely accurate frequency generation similar to cesium, although it suffers from the same pitfalls. Being only slightly less accurate than cesium, the rubidium standards have a high cost associated with them. They are not portable devices and must be in a fixed location. The rubidium standards also have the same depletion problem as cesium.

The choice for most consumer electronics is the quartz crystal oscillator. Quartz operates via the piezoelectric effect. Voltage applied to the crystal causes it to vibrate in a very stable and predictable way. A desired frequency of oscillation can be obtained by cutting and shaping the quartz. Quartz has many advantages. For example, it is inexpensive (compared to cesium and rubidium), it is small (crystals can be obtained in package sizes less than 3.2 × 2.5 mm) and has a high Q (> 500K for larger blanks). These traits have made quartz the industry choice for timing devices for many decades.

QUARTZ CRYSTAL OSCILLATORS

Crystal oscillators (XO) come in many types, shapes and sizes. XOs are comprised of a quartz crystal and driving circuitry. These bare bones clocks are inexpensive and small, but offer limited accuracy as the crystal will wander in frequency approximately ±30 ppm over temperature. Temperature-controlled crystal oscillators (TCXO) use a compensation voltage to correct for the crystal's natural temperature drift. This is accomplished with a classic thermistor resistor network or a polynomial generator. TCXOs are tighter stability over temperature (±0.25 ppm can be obtained) and are small (2.0 × 2.5 mm is available) with low power consumption (2 mA in some cases). Microprocessor controlled crystal oscillators (MCXO) use a microprocessor to correct for the crystal's natural temperature drift by sensing the temperature of operation and using that data to correct the frequency of the oscillator. These oscillators can achieve stabilities of ±0.1 ppm over temperature, but have a slightly larger footprint, consume more power and have degraded noise characteristics due to the microprocessor running in the oscillator. Oven-controlled crystal oscillators (OCXO) and double oven-controlled crystal oscillators (DOCXO) offer the greatest stability that quartz crystal oscillators have to offer. By heating the circuitry, the crystal stays at an almost constant temperature, nearly eliminating the natural temperature drift of the crystal. Stabilities of parts in 10^{-10} over temperature are achievable, but at the trade-off of footprint (at least 1 inch square) and power consumption (possibly over 1A).

Despite having a high Q, the quartz crystal oscillator is not perfect. Ideally a sinusoidal oscillator would produce a voltage such that

$$V(t) = V_o \cos(2\pi f_o t) \tag{1}$$

where

V_o = amplitude
f_o = frequency
t = time

However, real world oscillators have some amplitude fluctuations and phase fluctuations present within them and behave as in[4]

$$V(t) = [V_o + \lambda(t)]\cos(2\pi f_o t + \phi(t)) \tag{2}$$

where

$\lambda(t)$ = amplitude fluctuation
$\Phi(t)$= phase fluctuation

The frequency of the oscillator is affected by several factors; temperature, long-term drift and short-term instability. Quartz is quite sensitive to temperature variations, and the fre-

quency generally drifts between ±30 ppm over the temperature range of the crystal. Long-term drift, also called aging, is a natural phenomenon of the quartz and is well understood. The aging characteristic is defined as

$$f(t) = A\left(\ln\left(B_t + 1\right) + f_o\right) \qquad (3)$$

where

t = time in days
A, B and f_o = constants determined from least squares fit (as per MIL-PRF-55310D)[7]

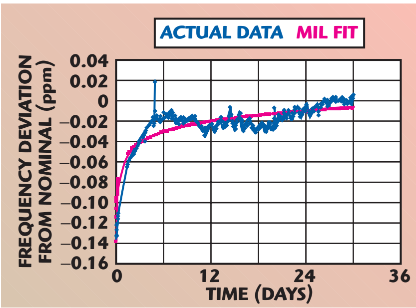
Figure 1 shows the aging characteristic of quartz is a natural log function that slows over time. This means that the frequency drift of the oscillator will diminish as time passes. This is a desirable effect in terms of long-term performance. The oscillator will drift but that rate of change will slow and the oscillator in effect will become more stable.

Short-term stability or Allan variance (AVAR) is a measurement of the short-term frequency variations from the oscillator. Generally AVAR is specified relative to a particular gate time. For example, a 20 ms gate time may be selected and 100 samples taken and applied to the following formula:

$$\text{AVAR} = \frac{1}{2(N-1)} \sum_{i=0}^{N} \left(f(i) - f(i-1)\right)^2 \qquad (4)$$

where f(i)-f(i-l) is the difference between successive frequency measurements (as per MIL-PRF-55310D).[7]

The result provides an approximation for how stable the oscillator is reading-to-reading at the given gate time. By lengthening the gate time, the oscillator's Allan variance decreases showing it is more stable over longer averaging periods.



▲ *Fig. 1 Typical TCXO aging data.*

JITTER

This measurement shows that the instantaneous frequency of the oscillator is not constant, but varies slightly about the nominal frequency creating an uncertainty in the frequency at any given point in time. This frequency change can be viewed as a change in time of the waveform edge from the ideal nominal frequency edge. This change in time of the edge is called jitter. ***Figure 2***[1] illustrates the jitter effect on a square wave.

Jitter can be measured in the time domain and is expressed in a peak-to-peak time variation of the edge. However, this method may not be very useful in some applications because the variations in the edges are coming from the entire band of frequencies and exaggerating the magnitude of the jitter. Most real world applications will operate within a certain band of frequencies and therefore the jitter effect only needs to be measured in that band. To effectively see and measure jitter over a particular frequency band, conversion to the frequency domain must take place.

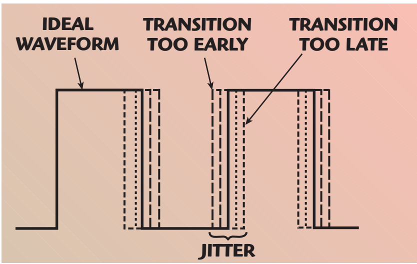
PHASE NOISE

This conversion to the frequency domain results in a measurement called phase noise. It is expressed as a graph of power vs. frequency. To understand this measurement the jitter-to-phase noise conversion must be explored. One way to measure jitter is to measure the variance of each period from the average period, such that[5]

$$\sigma_c^2 = \lim_{N \to \infty} \left(\frac{1}{N} \sum_{n=1}^{N} \left(\tau_n - \tau_{avg}\right) \right)^2 \qquad (5)$$

This RMS cycle jitter can then be used to calculate the phase noise at a given frequency, as

$$\mathscr{L}(f) = \frac{\sigma_c^2 f_{osc}^3}{f^2} \qquad (6)$$



▲ *Fig. 2 Time domain representation of jitter.*[1]

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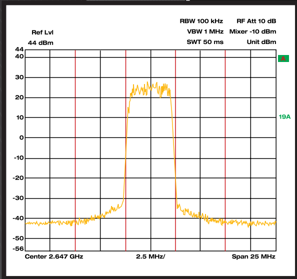

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where

f_{osc} = frequency of the oscillator
f = frequency offset from the carrier[5]

This computation can be done at many discrete frequencies and compiled into graphical form. Equations 5 and 6 assume that there is no l/f noise or burst noise.[5] In real world oscillators these noise sources are present.

If real world components and the noise that is generated in real world circuits is taken into account the phase noise calculation gets more complex. The Leeson equation as shown below provides an estimation for how circuit noise and circuit elements factor into the phase noise measurement.[6]

$$\mathcal{L}(f_m) = 10\log\left\{\frac{FkT}{2P_{avs}}\left[1+\frac{f_c}{f_m} + \left(\frac{f_o^2}{2f_m Q_l}\right)^2\left(1+\frac{f_c}{f_m}\right)\right]\right\} \quad (7)$$

where

Q = loaded Q of the circuit
f_n = frequency from the carrier
f_c = flicker noise corner frequency
f_o = carrier (oscillator) frequency
T = temperature in Kelvins
P_{avs} = power through the resonator

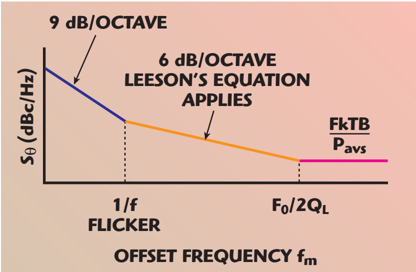


Fig. 3 Phase noise plot showing determining factors.[6]

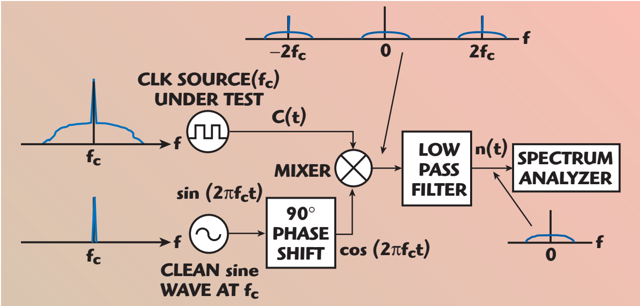


Fig. 4 Phase noise test setup.[2]

F = noise factor of the active device
k = Boltzmann constant

Figure 3 shows how this equation fits into the plot of phase noise. It can be seen that close into the carrier, flicker noise dominates the curve and has a cut-off frequency of the corner frequency of the active device. The middle portion of the phase noise plot follows Leeson's equation and is a combination of loaded Q, noise factor, power and temperature.[6] For frequencies above $f_o/(2Q_l)$, the floor is determined by noise factor, temperature and power.

From this plot it is fairly easy to come up with guidelines for minimizing phase noise in oscillator designs. Use devices with low flicker noise. Since the 9 dB/octave section is dominated by this quantity, reduction of circuit flicker noise is of great concern. BJTs have a much lower flicker noise than FETs making them more suitable for low phase noise applications. The 6 dB/octave section implies that a higher circuit Q is of great interest as is noise factor and power. OCXOs utilize crystals with higher Qs than TCXOs. Higher drive power is also desirable because that is the driving factor for the phase noise floor (frequencies above $f_o/(2Q_l)$). This comes with a tradeoff because higher drive levels usually result in some phase noise degradation close in to the carrier.

PHASE NOISE MEASUREMENT

Measuring phase noise is not an easy feat. Most spectrum analyzers do not have the resolution to measure the phase noise of a crystal oscillator directly. ***Figure 4*** shows the normal configuration for phase noise testing. An 'ideal' source that is the same frequency as the oscillator under test is mixed against the oscillator. This produces the products of these two signals and also the difference. Using a low pass filter, the products are stripped away leaving only the difference, which will be the oscillator noise, if the 'ideal' source is exactly the same frequency as the oscillator.

This resulting noise signal can then be measured by a spectrum analyzer and displayed as a graph of dBc/Hz vs. Hz. This graph is displaying in terms of power how far the noise (at each particular frequency) is below the carrier (oscillator's desired signal). ***Figure 5*** shows an OCXO (two-inch square package) plot that has a floor of around –170 dBc and a 10 Hz performance of approximately –130 dBc. ***Figure 6*** depicts the phase noise performance of a 5 mm × 7 mm TCXO, which has a floor of around –155 dBc and a 10 Hz performance of approximately –90 dBc. Both plots are typical phase noise measurement plots.

OCXO phase noise performance is commonly much better than that of TCXOs. This is due to several factors. The OCXO is inherently more stable than the TCXO giving it improved phase noise performance. This is because the crystal in the OCXO is thermally stable where the TCXOs crystal is responding to the environmental changes. Even though there is compensation for those ambient temperature changes it is nowhere near as stable as the OCXO. Secondly, OCXOs use different crystal cuts than TCXOs and as a result have a higher Q. A TCXOs crystal Q is approximately 30,000 to 40,000, where an SC-cut ovenized crystal may have a Q approaching 1,000,000. This higher Q directly improves phase noise in the oscillator. Lastly, TCXOs are designed to be much smaller and have very low power consumption. This means most of them are constructed using FETs instead of BJTs. FETs have a much higher inherent flicker noise, which worsens the phase noise characteristics of the oscillator. Ovenized units designed for stability use BJTs, which have lower noise specifications and in turn better phase noise performance.

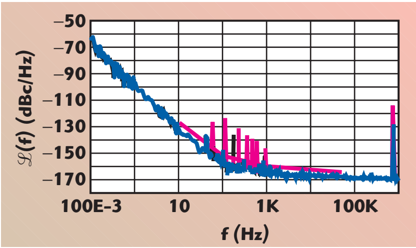


Fig. 5 Greenray Industries YH1321-4 OCXO phase noise performance.

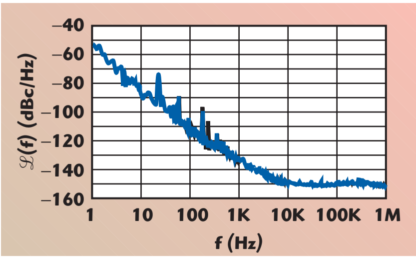


Fig. 6 Greenray Industries T75 TCXO phase noise performance.

MEASUREMENT FLAWS

One of the issues with this measurement system is that the oscillator is mixing under test against an 'ideal noise free' oscillator. Since there is no such real world device the oscillator must be mixed under test against something known to be much better than the unit under test and assume that the 'reference' injects no noise. Another option is to mix it against something that is equivalent to the unit under test, and assume equal contribution from both oscillators. Both of these methods are inherently flawed. By mixing against something better one is assuming no degradation from the reference. If a reference that is 10 times better is available, the effects are minimal, but there is still degradation in the measurement. By mixing against an equal source and assuming equal contribution, 3 dB is added to the target, which allows the operator to accept results that really exceed the intended specification because there is another noise source. However, how does one really know that they are equal? In fact, most times they are not equal. One source will have better

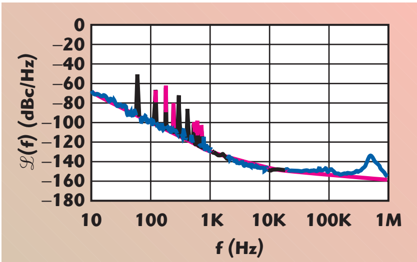


Fig. 7 TCXO with noisy supply.

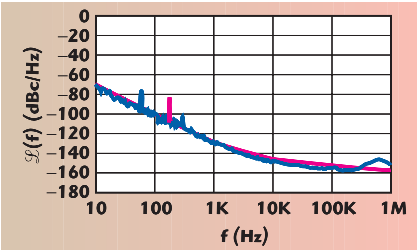


Fig. 8 TCXO with cleaner supply.

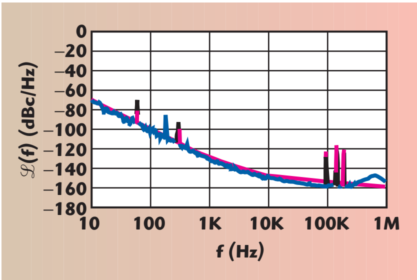
noise characteristics than the other and this creates ambiguous data. Actually, if the unit picked as a 'reference' ends up being a relatively clean source and equal contribution is assumed, the operator could be accepting results that are in reality failing, but he or she has no way of knowing.

OUTSIDE NOISE/INTERFERENCE

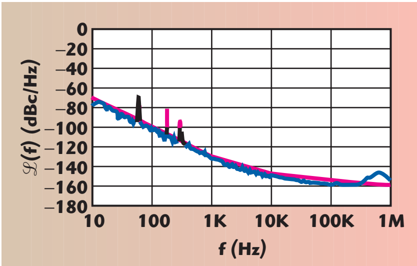
Aside from the circuit characteristics, measurement assumptions and random crystal fluctuations, the oscillator and its phase noise behavior are affected greatly by the power supplied to the oscillator. ***Figure 7*** shows a TCXO with a noisy power supply. Spurs can be seen between 80 and 1000 Hz with an amplitude of approximately –40 dBc. ***Figure 8*** shows the same TCXO tested with a cleaner supply and only two spurs remain with amplitudes of approximately –20 dBc. The oscillator manufacturer has no control over what power supply the operator uses in their application and therefore just the oscillator's phase noise performance is characterized and not the impact of the supply on the oscillator. As a result, the cleanest possible supplies are used for these measurements.

Nearby interference can also show up in phase noise measurements. ***Figure 9*** shows the TCXO with a clean supply running it, but a computer monitor running about three feet away from the oscillator and test set. ***Figure 10*** shows the same test with the monitor turned off. Again, the spurs appearing in ***Figure 9*** are not from the oscillator and are therefore eliminated by shutting off the computer monitor during testing.

These real world plots show where the disconnect between consumer and manufacturer comes into play. Oscillator manufacturers attempt to characterize just the oscillator's phase noise performance, and as a result go to great lengths to accomplish this. Clean power supplies or even batteries are used in an attempt to reduce the noise injected from the outside, reducing or eliminating any nearby interference such as computer monitors or other equipment running in the immediate area. Some manufacturers even build a Faraday cage, or shielded room to reduce outside interference. Many use baffles or buffers to reduce airflow around the oscillator making it more thermally



▲ *Fig. 9 TCXO with monitor running.*



▲ *Fig. 10 TCXO with monitor off.*

stable. What is wrong with that? The result of all of these efforts is an attempt to eliminate all outside interferences so just the oscillator's performance can be measured. However, customers purchasing the oscillators may not realize that these great lengths were taken and the impact of the external effects. Secondly, the oscillator is not going to be running in a shielded room all by itself. It will be in a larger assembly with temperature gradients and non-ideal power supplies and probably other machinery or computers running nearby. Therefore, if a user looks at a phase noise plot and assumes this is what they have in their system there could be a substantial descrepancy. If no headroom is allowed for in their design, serious problems could occur.

Lastly, there are two other factors that directly affect the phase noise performance that the oscillator manufacturer has no control over. Vibration affects the phase noise of crystal oscillators, as described in the following formula:

$$\mathscr{L}(f_v) = 20\log\left(\frac{\Gamma A f_o}{2 f_v}\right) \qquad (8)$$

where

Γ = oscillator's g-sensitivity
A = amount of gs in the vibration
f_o = frequency of the oscillator
f_v = frequency of the vibration[10]

Most military designers are well aware of this degradation and work with the manufacturer to understand and characterize the oscillator under the vibration it will see in the field.

Another factor that hinders phase noise performance is multiplication. Many users will purchase an oscillator in the megahertz region and multiply it up into the gigahertz range. This is acceptable and will provide a stable oscillator but its phase noise is reduced by

$$\mathscr{L}(f) = 20\log(N) \qquad (9)$$

where

N = factor of multiplication[10]

Vibration and multiplication are two processes that the user might be subjecting the oscillator to that the manufacturer would have no idea about and cannot give the user an indication about performance degradation. Therefore, if at all possible, it is important for designers to understand the affect these variables have on the time base and to work with the oscillator manufacturer so that the required quality of the timing device is appropriate for a particular application.

CONCLUSION

Phase noise is a very important aspect of any real world time base. From the previous discussion an oscillator designer can come to several key conclusions that will help minimize the phase noise in an oscillator design. Maximizing Q, using low noise active devices and increasing resonator power are the three main things to be done when reducing the phase noise in an oscillator. The oscillator designer can also see that the method of testing greatly affects the results of the test. Low noise supplies

or batteries can be used to reduce phase noise from supply ripple. Nearby interference from computers, lighting, or other machinery should be eliminated. Finally, thermal buffers can be used to ensure that the test is only measuring the oscillator's noise and not noise from outside entities.

Understanding the phase noise phenomenon and characterizing its effects are paramount to communication system design. From a user's perspective this article should be helpful as a tool to understanding phase noise better and also understanding how it is tested and what factors outside of the oscillator will degrade its performance. There have been many instances where users did not specify what was required, or did not obtain the performance that they thought they were getting. This is a result of how the oscillator industry goes about performing the phase noise test. The attempt is made to eliminate any and all outside influences on the oscillator so that the oscillator's noise characteristics can be assessed. This is legitimate and understandable, but the consumer of this product needs to be aware that the device will be used in less than ideal conditions and certainly more interference is present than was present when the phase noise was measured by the manufacturer. As noted above, other factors in the end application can cause phase noise issues that the user did not anticipate such as vibration and multiplication of the signal. At the very least system designers should communicate with the oscillator manufacturer to let them know what is the likelihood of noise in their system so that the oscillator manufacturer can relate a more realistic picture of the phase noise performance of the oscillator that is being purchased. ■

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John Esterline *holds a BSEE degree and is currently pursuing his master's of engineering in electrical engineering degree from Pennsylvania State University. He is currently an oscillator design engineer at Greenray Industries Inc. He has been involved in the testing and design of precision crystal oscillators for five years. He may be contacted via e-mail at jesterline@greenrayindustries.com.*

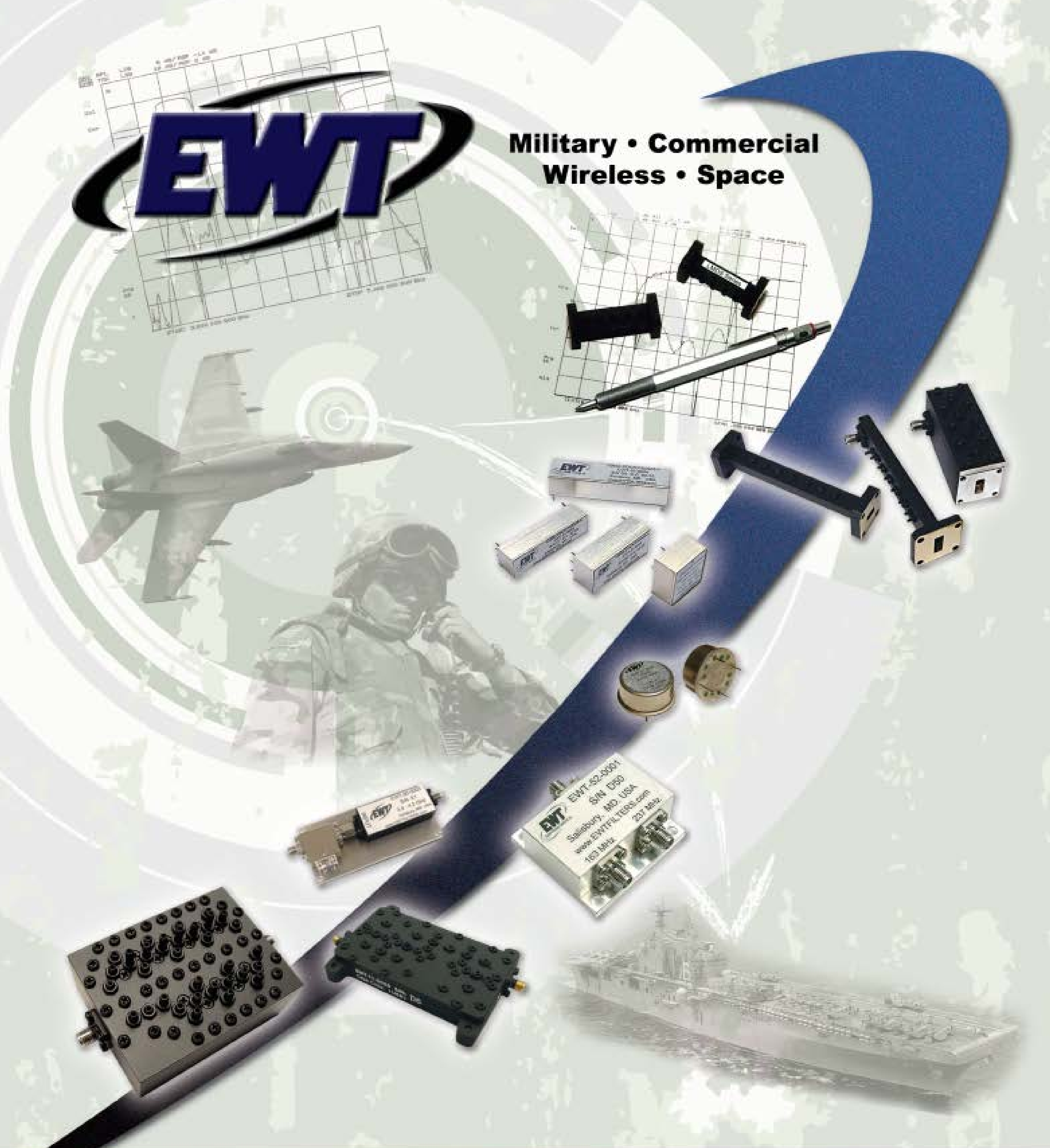
A NOVEL N-WAY DISTRIBUTED DOHERTY AMPLIFIER WITH IMPROVED EFFICIENCY AT HIGH PAR SIGNALS

A novel three-way distributed Doherty power amplifier, with improved efficiency at high peak-to-average ratio (PAR) signals such as W-CDMA and OFDM, is presented. This distributed Doherty amplifier consists of one main amplifier and two peaking amplifiers. In order to improve the efficiency at a high back-off power, the peaking amplifier structure is based on a dual-fed distributed amplifier form. The results of the three-way distributed Doherty amplifier, measured at 2140 MHz, produced a power gain of 11 dB. At that point, a power-added efficiency (PAE) of 39.5 percent was measured for a 9.5 dB PAR signal. A high linearity performance of –54 dBc ACPR is achieved after using memory-compensated digital predistortion.

In modern digital wireless communication systems, such as IS-95, PCS, W-CDMA, OFDM and so on, power amplifiers have progressed towards having a wide bandwidth and a large number of carriers. Recently, orthogonal frequency division multiplexing (OFDM) modulation has become an attractive technique for transmitting information efficiently within a limited bandwidth like WiBRO and WiMAX. However, since the OFDM signal consists of a number of independently modulated sub-carriers, it produces a higher peak-to-average power ratio (PAR) signal. A typical PAR for a 64-subcarrier OFDM signal is approximately 8 to 13 dB. When the number of sub-carriers is increased to 2048, the PAR is also increased to 11 to 16 dB. The available efficiency of the power amplifiers designed for these high PARs will be significantly deteriorated.

The Doherty amplifier is known as a technique for improving the efficiency at high output back-off power. Its primary advantage is the ease of configuration when applied to high power amplifiers, unlike other efficiency enhancement amplifiers or techniques such as switching mode amplifiers, EER, LINC and so on.[1] Recent results have been reported on

KYOUNG-JOON CHO AND WAN-JONG KIM
Dali Wireless
Sunnyvale, CA
JI-YEON KIM AND JONG-HEON KIM
Kwangwoon University
Seoul, Korea
SHAWN P. STAPLETON
Simon Fraser University
Burnaby, BC, Canada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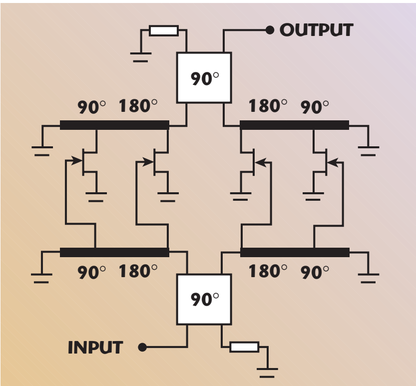
its use as a symmetric Doherty structure,[2] an asymmetric Doherty structure with uneven power transistors[3] and as an N-way Doherty structure using multi-paralleled transistors.[4] In the case of the symmetric Doherty amplifier, the maximum efficiency point is obtained at a 6 dB back-off power. The asymmetric Doherty amplifier can obtain a high efficiency at various back-off powers using a combination of uneven power device sizes for the main and peaking amplifier. Unfortunately, it is difficult to optimize the gain and output power of the asymmetric Doherty amplifier because of the different device matching circuits and the delay mismatch between the main and peaking amplifier. The conventional N-way Doherty amplifier has an efficiency enhancement over a conventional two-way Doherty structure by using multiple parallel transistors of identical devices. Its one drawback is that the total gain will be reduced due to the loss of the N-way input power splitter. Under low gain situations, this will increase the power dissipation of the driving amplifier.

In order to prevent the total gain of the Doherty amplifier from degrading, the peaking amplifier is designed using a dual-fed distributed amplifier structure. Distributed amplification is a technique whereby the power combining is performed directly at the transistor level without the need for an N-way power combiner.[5] Therefore, it is an easy configuration for combining peaking amplifiers.

OPERATIONAL PRINCIPLES OF AN N-WAY DISTRIBUTED DOHERTY POWER AMPLIFIER

The distributed power amplifier, which was described by Eccleston,[5] is a simple and feasible way to increase the gain performance of a power amplifier. The balanced dual-fed distributed amplifier approach shown in ***Figure 1*** can be formed by using a pair of identical single-ended distributed power amplifiers and quarter-wave hybrid couplers. As viewed on the single-ended distributed structures, a hybrid is used to feed both ends of the gate lines and another hybrid combines the waves appearing at the ends of the drain lines, thereby increasing the gain. The FET drain voltages and currents are equal when the FETs are spaced by a half-wavelength at the center frequency when using an input and output power splitter/combiner and an even number of FETs.

Figure 2 is the schematic of a new N-way distributed Doherty power amplifier, based on a distributed structure. The distributed Doherty amplification approach uses a power combiner and a splitter at the output and the input, respectively. This will improve the isolation between the main amplifier and the peaking amplifier. Both the N-main amplifiers and N-peaking amplifiers are combined using half-wave and quarter-wave microstrip lines.



▲ *Fig. 1 Diagram of a conventional balanced dual-fed distributed amplifier.*

In order to understand the load impedance variation of the N-way Doherty amplifier, it can be analyzed by the equivalent conversion of AC multi-current sources. ***Figure 3*** shows the operational principle of load modulation for the N-way distributed Doherty power amplifier. Im,M and Ip,P are ideal AC current sources for the carrier amplifiers and peaking amplifiers, respectively. The –90° phase difference between the main source and the peaking source is selected to compensate for the phase shift by the quarter-wave transformer, which is necessary to the Doherty operation. The output node voltage, V_0, through the load resistor, $R_0/(M+P)$, can be given by

$$V_0 = R_0(I'_M + I_P)/(M + P) = R_0(I_M\angle 90° + I_P\angle{-90°})/(M+P) \quad (1)$$

Thus, the impedance of the main amplifier, which can be seen from the quarter-wave transformer, Z'_m, can be written as

$$Z'_m = \frac{V_o}{I'_M} = \frac{R_o\left(I_M + I_P\right)}{\left(M+P\right)I_M} \quad (2)$$

Then, assuming that I_M and I_P are MI_m and PI_p, respectively, the impedance of the main amplifier, which can be seen from the main current souce, Z_m, can be found by

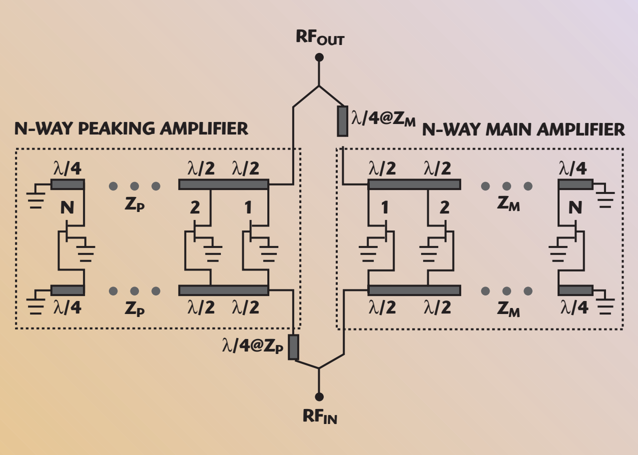
$$Z_m = \frac{R_o^2}{Z'_m} = \frac{\left(M+P\right)R_o I_m}{MI_P + PI_P} \quad (3)$$

If identical devices are used in the main and peaking amplifiers, the load impedance of an N-way Doherty amplifier could be modulated as a function of the number of N-ways. Finally, the load impedance of the N-way Doherty amplifier as a function of three operational regions of the peaking amplifier could be divided by:

- When I_p is 0, the load impedance becomes (1+P/M) Ro.
- When I_p is equal to I_m, the load impedance becomes R_o.
- When I_p is smaller than I_m, the load impedance transits from $N^* R_o$ to R_o.

The efficiency of an N-way Doherty amplifier, which is comprised of M, class B biased, main amplifiers and P, class B biased, peaking amplifiers, can also be divided into three input power regions (low, medium and peak levels).[4]

$$\eta = \begin{cases} \dfrac{\pi}{4}, \ v_o = v_{max} & \\[2ex] \dfrac{\pi}{4}\dfrac{\left(\dfrac{P}{M}+1\right)\left(\dfrac{v_o}{v_{max}}\right)^2}{\left(\dfrac{P}{M}+2\right)\left(\dfrac{v_o}{v_{max}}\right)-1}, & \dfrac{v_{max}}{\left(\dfrac{P}{M}+1\right)} \le v_o \le v_{max} \\[2ex] \dfrac{\pi}{4}\left(\dfrac{P}{M}+1\right)\dfrac{v_o}{v_{max}}, & 0 \le v_o \le \dfrac{v_{max}}{\left(\dfrac{P}{M}+1\right)} \end{cases} \quad (4)$$

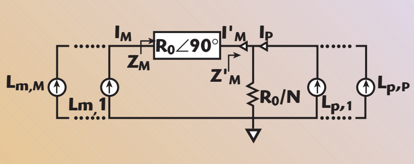


▲ *Fig. 2 Schematic of a new N-way distributed Doherty amplifier.*

where

v_o = output voltage for a given power level
v_{max} = maximum output voltage for a given power level

Figure 4 shows the calculated efficiency of the N-way Doherty amplifier, for different topologies, at various back-off powers: 3.5, 7.9, 9.5 and 12 dB. The desired location of the peaking point of an N-way Doherty amplifier can be easily derived from Equation 4. This peaking point is defined as the extended back-off state, X_{BO}, and is given

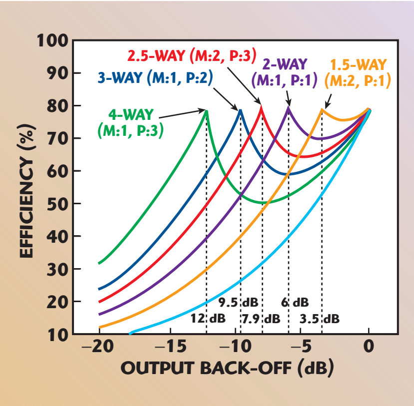


▲ *Fig. 3 Load impedance variation of the N-way distributed Doherty amplifier.*

$$X_{BO} = 20\log_{10}\left(\frac{P}{M}+1\right)\ [\text{dB}] \qquad (5)$$

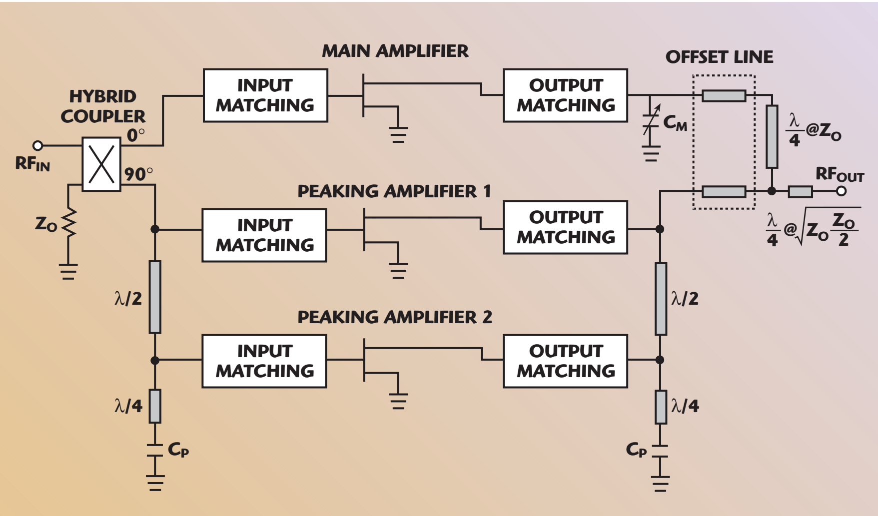
where

P = number of peaking amplifiers
M = number of main amplifiers



▲ *Fig. 4 Calculated efficiency of the N-way Doherty amplifier at various back-off powers.*

The schematic of the proposed three-way distributed Doherty amplifier is shown in ***Figure 5***. In order to provide high efficiency at a back-off power of 9.5 dB, two peaking amplifiers are combined using the dual-fed distributed structure. It consists of half-wave and short-circuited quarter-wavelength microstrip lines. This structure should insure that the total gain of the N-way distributed Doherty amplifier should not be degraded



▲ *Fig. 5 Schematic of the proposed three-way distributed Doherty amplifier.*

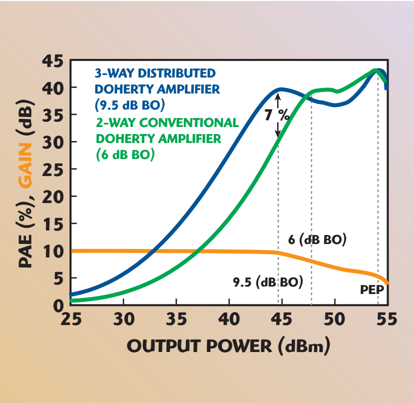
compared to the conventional N-way Doherty amplifier.

DESIGN AND SIMULATION

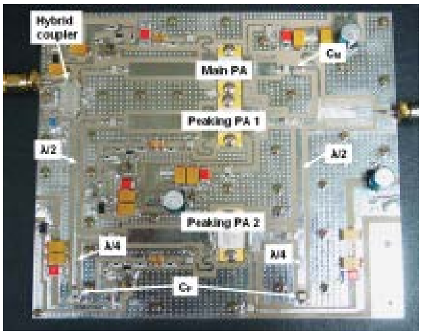
The proposed three-way distributed Doherty amplifier design uses the same device for the one main amplifier and the two peaking amplifiers. The main amplifier is biased in the class AB mode and the peaking amplifiers are biased in the class C mode. The devices are Freescale high power MRF21045 LDMOS FETs with a peak envelope power (PEP) of 45 W. The simulation for the Doherty circuit was performed using Agilent's ADS software.

Optimization of the N-way distributed Doherty amplifier required the following:

- Evaluating the optimum number of main amplifiers (M) and peaking amplifiers (P) for the given back-off power
- Bias adjustment of the main amplifier in class AB mode
- Gate bias adjustment of the peaking amplifiers in class C mode
- Offset line optimization for achieving a peaking point at the desired back-off power



▲ *Fig. 6 Simulated PAE and gain performance of the proposed three-way distributed Doherty amplifier.*



▲ *Fig. 7 The three-way distributed Doherty amplifier.*

The shunt capacitors, C_P, are inserted so that both gain and efficiency of the Doherty amplifier are optimized. The other shunt capacitor, C_M, is also used for optimizing the linearity.[6]

Figure 6 shows the simulated PAE and gain performance of the proposed three-way distributed Doherty amplifier using a single tone at 2140 MHz. The operating point of the class AB biased main amplifier is: $I_{DQ} = 510$ mA, $V_{GS} = 3.82$ V and $V_{DS} = 27$ V. The operating points of the class C biased peaking amplifiers are:

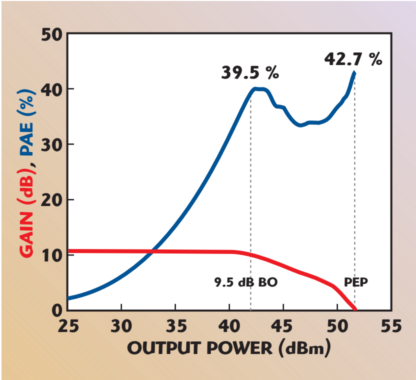
Peaking amplifier 1:

$I_{DQ} = 0$ mA, $V_{GS} = 2.4$ V and $V_{DSc} = 27$ V.

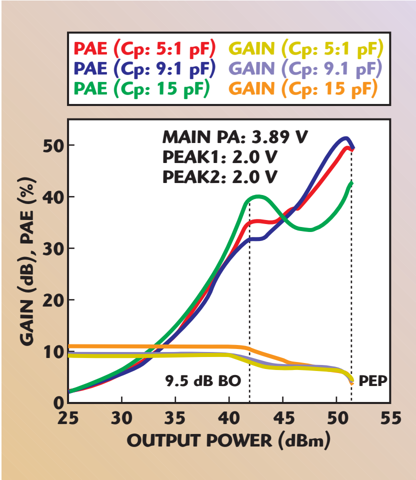
Peaking amplifier 2:

$I_{DQ} = 0$ mA, $V_{GS} = 2.6$ V and $V_{DS} = 27$ V.

The output impedance of the combined peaking amplifier using a dual-fed distributed structure was (4.65+j 2.1)Ω. The offset line is necessary to prevent leakage power be-



▲ *Fig. 8 Measured PAE and gain performance of the proposed three-way distributed Doherty amplifier.*



▲ *Fig. 9 Measured PAE and gain as a function of the shunt capacitance C_M.*

tween the main amplifier and the peaking amplifier. An offset line of approximately 0.25 λ was inserted; this corresponds to an optimum output resistance of 521 Ω. From the simulated results, a 43 percent PAE was obtained at a PEP of approximately 200 W. Consequently, a 40 percent PAE at 9.5 dB back-off power from the peak efficiency point was achieved. This was an efficiency improvement of approximately 7 percent compared to that of the two-way conventional Doherty amplifier at a 6 dB peaking point. A gain of approximately 10.5 dB was obtained from 2130 to 2150 MHz. This was similar to that of the conventional two-way Doherty amplifier.

FABRICATION AND MEASUREMENTS

A new three-way distributed Doherty amplifier using Freescale's MRF21045 was fabricated and measured at a center frequency of 2140 MHz and a single carrier W-CDMA signal with a peak-to-average ratio of 9.8 dB. ***Figure 7*** is a photograph of the three-way distributed Doherty power amplifier. ***Figure 8*** shows the measured PAE and gain performance of the proposed three-way distributed Doherty amplifier, using a single-tone, 2140 MHz signal. The main amplifier's operating point is: I_{DQ}= 480 mA, V_{GS}= 3.9 V. The operating points of the peaking amplifiers are:

Peaking amplifier 1: I_{DQ} = 0 mA, V_{GS} = 2.1 V.

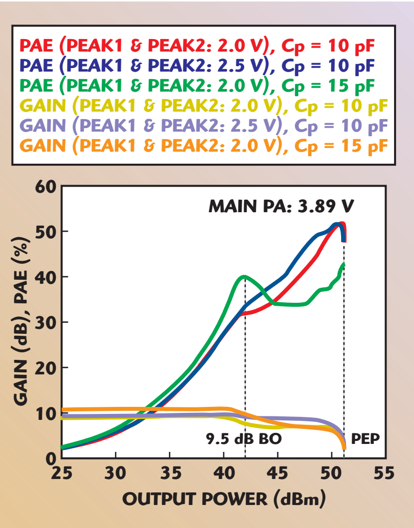
Peaking amplifier 2: I_{DQ} = 0 mA, V_{GS} = 1.9 V.

Shunt capacitors, C_P = 15 pF and C_M = 0.5 pF, are used. A 42.7 percent PAE at a PEP of 131 W and a 39.5 percent PAE at 9.5 dB back-off are achieved. A gain of approximately 11 dB was obtained at 9.5 dB back-off.

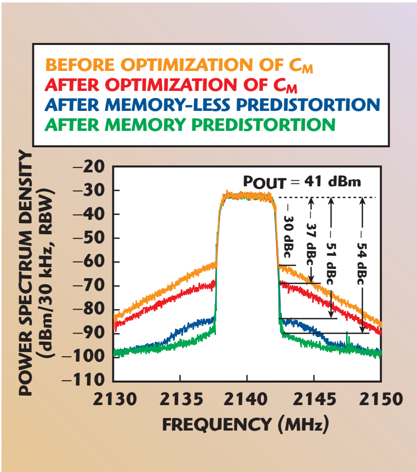
Figure 9 shows the measured PAE and gain as a function of the shunt capacitor CP values of 5.1, 9.1 and 15 pF, in order to find the optimal capacitance that could affect the gain and PAE of the Doherty amplifier. It was tested with a single-tone, 2140 MHz signal and the gate biases of the main amplifier and peaking amplifiers are fixed to 3.89 and 2.0 V, respectively.

Figure 10 shows the measured PAE and gain variations as a function of the shunt capacitor, C_P, and the bias of the peaking amplifiers using a single-tone, 2140 MHz signal. Optimization of C_P and of the bias point of the two peaking amplifiers produced efficiency and gain improvement of approximately 8 percent and 2 dB at 9.5 dB back-off, even though the PAE is reduced at PEP. Consequently, a high PAE of the distributed Doherty amplifier could be obtained by adjusting the shunt capacitor, C_P, and the bias of the peaking amplifiers.

Figure 11 shows the measured W-CDMA, one-carrier spectrum of the three-way distributed Doherty amplifier before and after optimization of C_M, and after digital predistortion. The test bench set-up of the digitally pre-distorted power amplifier and the operation princi-



▲ *Fig. 10 Measured PAE and gain variation as a function of Cp and peaking amplifier's bias.*



▲ *Fig. 11 Measured WCDMA one-carrier spectrum of the three-way Doherty amplifier.*

TABLE I

RESULTS FOR THE PROPOSED THREE-WAY DISTRIBUTED DOHERTY AMPLIFIER

Contents	Three-way Distributed Doherty PA		Three-way Distributed Doherty PA Using Digital Predistortion	
	Before Opt. (C_M = 0 pF)	After Opt. (C_M = 0.5 pF)	Memoryless (7th Order)	Memory (7th Order + 2 Terms)
ACPR (dBc) (@2.5 MHz offset)	–30	–37	–51	–54
Operating Points	Main PA: VGS=3.9 V Peak PA1: VGS=2.1 V Peak PA2: VGS=1.9 V CP=15 pF	Main PA: VGS=3.9 V Peak PA1: VGS=2.5 V Peak PA2: VGS=2.4 V CP=10 pF	Main PA: VGS=3.79 V Peak PA1: VGS=3.1 V Peak PA2: VGS=2.5 V CM=9.1 pF, CP=0.5 pF	
PAE (%) (10 dB back-off)	39.5	35	33	

ples have been published previously.[8] After optimization of C_M, a linearity improvement of 7 dB was achieved at a +2.5 MHz offset frequency. The operating points were V_{GS} =3.79 V (Main PA), V_{GS} = 3.1 V (Peaking PA1) and V_{GS} = 2. 5 V (Peaking PA2). Shunt capacitors, C_P = 9.1 pF and C_M = 0.5 pF, were used. In order to achieve high linearity, both memory-less and memory-based digital predistortions were applied. The ACLR performance of –51 dBc after memory-less and –54 dBc after memory compensation were obtained at 41 dBm output power and +2.5 MHz offset frequency.

Table 1 shows the summary results of the proposed three-way distributed Doherty power amplifier. This research confirms that both high efficiency and high linearity for W-CDMA and OFDM power amplifiers can be achieved.

CONCLUSION

A novel Nth-way distributed Doherty amplifier was proposed to obtain high efficiency at various back-off power states. The peaking amplifiers of the N-way distributed Doherty amplifier were combined using a dual-fed distributed structure. From the measured results, the three-way distributed Doherty amplifier yielded a 39.5 percent PAE at 9.5 dB back-off power and a power gain of 11 dB. After final optimization for linearity and efficiency, –54 dBc ACPR and 33 percent PAE are achieved, using memory compensated digital predistortion. ■

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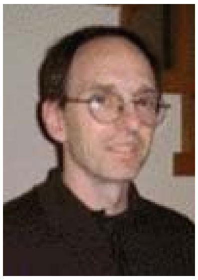
Kyoung-Joon Cho *received his BS degree in information and communication engineering from Anyang University, Anyang, Korea, in 1998, and his MS and PhD degrees in radio science and engineering from Kwangwoon University, Seoul, Korea, in 2000 and 2004, respectively. He was a postdoctoral fellow in the school of engineering science at Simon Fraser University, British Columbia, Canada, from 2004 to 2007. He is currently a senior HPA designer at Dali Wireless, Sunnyvale, CA. His research interests include MMIC/hybrid high efficient power amplifier design and linearization techniques.*

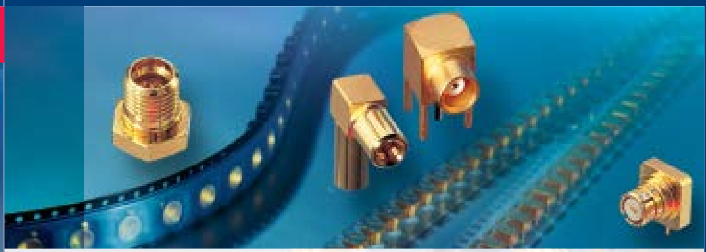
Wan-Jong Kim *received his BS and MS degrees in radio science and engineering from Kwangwoon University, Seoul, Korea, in 1999 and 2001, respectively, and his PhD degree from the school of engineering science at Simon Fraser University, British Columbia, Canada, in 2006. Since 2007, he has been working as a senior DSP designer at Dali Wireless, Sunnyvale, CA. His research interests include peak-to-average power ratio reduction techniques, RF/DSP integrated system and digital linearization techniques.*

Ji-Yeon Kim *received her BS and MS degrees in radio science and engineering from Kwangwoon University, Seoul, Korea, in 2002 and 2004, respectively. She is currently working toward her PhD degree in radio science and engineering, Kwangwoon University, Seoul, Korea. Her research interests include high efficiency power amplifier design and linearization techniques.*

Jong-Heon Kim *received his BS degree in electronic communication engineering from Kwangwoon University, Seoul, Korea, in 1984, his MS degree in electronic engineering from Ruhr University, Bochum, Germany, in 1990, and his PhD degree in electronic engineering from Dortmund University, Dortmund, Germany, in 1994. Since March 1995, he has been a professor in the department of radio science and engineering at Kwangwoon University, Seoul, Korea. His current interests include digital linearization of power amplifiers and transmitters, smart power amplifiers and integrated RF/DSP design.*



Shawn P. Stapleton *received his BS, MS and PhD degrees in engineering from Carleton University, Ottawa, Ontario, Canada, in 1982, 1984 and 1988, respectively. Since 1988, he has been a professor in the school of engineering science at Simon Fraser University, British Columbia, Canada. His research interests include integrated RF/DSP applications for wireless communications, GaAs MMIC circuits and power amplifier linearization. He has developed a number of adaptive linearization techniques ranging from feedforward, active biasing, work function predistortion to digital baseband predistorters.*



A HIGH EFFICIENCY AND GAIN DOHERTY AMPLIFIER FOR WIRELESS MOBILE BASE STATIONS

This article presents a high gain and efficiency Doherty amplifier for WLAN mobile base stations with three advanced methods. First, a Doherty amplifier is developed with uneven power drive, which provides more input power to the peak amplifier for full power operation and appropriate load modulation. Second, a special inverted Doherty topology is proposed in order to optimize the average efficiency of the Doherty amplifier. Finally, a three-stage structure is adopted to increase the total gain. These methods are applied to implement a Doherty power amplifier using Freescale MRF6S21050L devices. The power-added efficiency (PAE) and adjacent channel leakage ratio (ACLR) are 33 percent and –42 dBc, which represents an improvement of 3.2 percent and 2 dB, respectively, compared with a conventional Doherty power amplifier and its gain reaches 45 dB by adopting a three-stage structure.

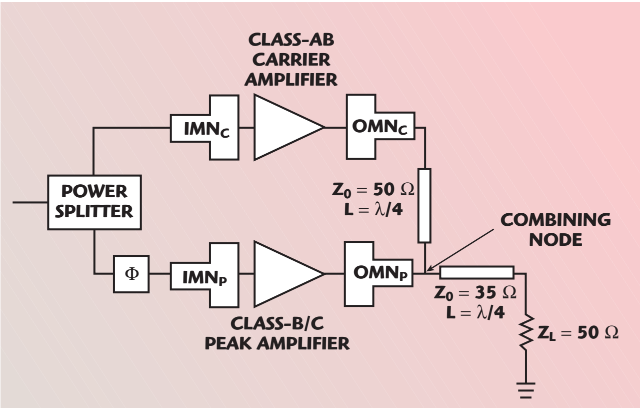
Wireless communications have recently progressed to require an increase in bandwidth and the number of carriers for high data rate capability. In addition to the wide bandwidth, the instantaneous transmit power of the wireless communication systems, such as wideband code division multiple access (WCDMA) or orthogonal frequency division multiplex (OFDM), vary widely and rapidly, carrying high peak-to-average ratio (PAR) signals.[1] The mobile wireless base station power amplifiers for the systems require a high linearity to amplify the high PAR signal source without distortion.[2,3] Another requirement of power amplifiers for mobile wireless communications systems is high efficiency.[4] The communications systems are reduced in both size and cost but require a power amplifier with high efficiency.[5–7]

In order to accomplish these requirements, the techniques that can improve the linearity and efficiency of the mobile base station power amplifier and overcome the wideband effect are hot issues in the research community. As for conventional techniques, the simplest method

X.Q. CHEN, Y.C. GUO AND X.W. SHI
Xidian University
Xi'an, China

is to back-off signals from the saturation region to the linear region at the cost of reduced efficiency of the power amplifier. Class-A, -B, or -AB may be used with a low efficiency from 12 to 20 percent, due to the output back-off.[8] Other methods involve predistortion or elimination and restoration techniques or feedforward.[9–11] However, these techniques need additional components that result in an increase in cost, size and power dissipation.[12] In order to solve these problems, a Doherty amplifier is the most promising candidate for the application, as shown in ***Figure 1***.

The fundamental operation theory has been well described in previous literature. It has high linearity and efficiency for wideband signals and has been studied extensively for the application, as depicted in ***Figure 2***. However, the conventional Doherty power amplifier has its limitation. Because of its lower bias point, the current level of the peaking cell is always lower than that of the carrier. The load impedances of both cells cannot be fully modulated at the value of the optimum impedance for a high power match. Thus, neither cell can generate full output power. In this article, two advanced methods that represent good approaches to solve these problems are presented. First, when the magnitude of the instantaneous input signal increases, the gate bias voltage of the peak cell increases according to the signal's envelope. Therefore, uneven power drive is possible to achieve a good Doherty operation. Second, the inverted Doherty topology is another efficient method to improve the carrier cell efficiency at low levels when the peak cell is off. The implementation of the amplifier is simple and results show excellent performance.

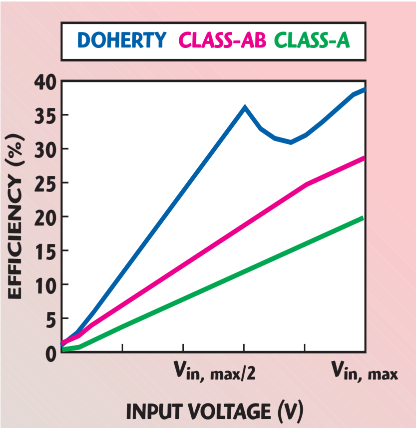


▲ *Fig. 1 The classical Doherty power amplifier.*

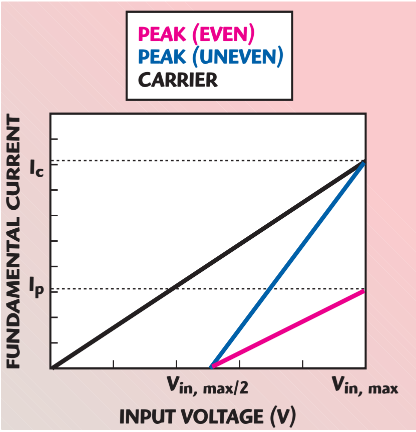
ADVANCE DESIGN METHODS OF A DOHERTY POWER AMPLIFIER

Uneven Power Drive

Figure 3 depicts the different fundamental current characteristics of the Doherty power amplifier with even drive and uneven drive. Usually, in even drive, the load impedances of the carrier amplifier and the peak amplifier cannot be completely modulated at the optimum impedance and the peak amplifier cannot generate full power because it is biased in class-C, and its current Ip remains at a low level.[13] The linearity of the even drive amplifier is more complex than that of a class-AB amplifier.



▲ *Fig. 2 Efficiencies of the Doherty, class-A and class-AB amplifiers.*



▲ *Fig. 3 The Doherty amplifier's fundamental current vs. input voltage.*

Figure 4 shows the load impedance of both amplifiers versus input voltage.

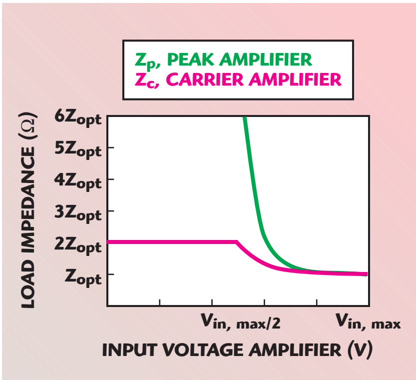
$$
Zp = \begin{cases} \dfrac{Z_T^2}{Z_L}, & -0 < V_{in} < V_{in,max/2} \\ Z_L\left(1+\dfrac{I_p}{I_c}\right), & -V_{in,max/2} < V_{in} < V_{in,max} \end{cases} \tag{1}
$$

$$
Zp = \begin{cases} \infty, & -0 < V_{in} < V_{in,max/2} \\ Z_L\left(1+\dfrac{I_p}{I_c}\right), & -V_{in,max/2} < V_{in} < V_{in,max} \end{cases} \tag{2}
$$

where

Z_L = load impedance of the Doherty amplifier
I_C = fundamental currents of the carrier amplifiers
I_P = fundamental currents of the peak amplifiers
Z_C = output load impedances of the carrier amplifiers
Z_P = output load impedances of the peak amplifiers

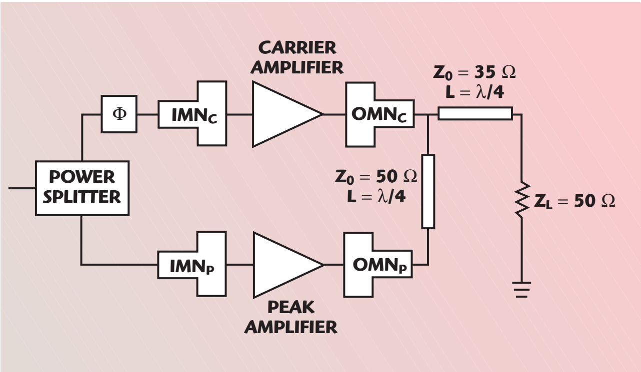
In the low power region, the linearity of the amplifier is entirely determined by the carrier cell. Therefore, the carrier cell should be highly linear for its optimized load impedance. In the high power region, the current level of the peaking cell plays an important role in determining the load modulation of the amplifier. For the asymmetric amplifier with even power drive, the fundamental current of the peaking cell is insufficient to achieve the full load modulation. The load impedances of both cells are larger



▲ *Fig. 4 Load impedance vs. input drive.*

than the optimum values in the high-power region. As a result, the carrier and peaking cells are driven into saturation without producing full power. Thus, the amplifier's linearity is seriously affected, as is its power level.

In order to enhance the output power from the peak amplifier, a Doherty amplifier with uneven power drive is proposed, applying more power to the peak cell. Since the amplifier now has an uneven power drive, the linearity of the amplifier is improved due to proper power operation without severe saturation. The linearity is further enhanced by the harmonic cancellation of the two cells at appropriate gate biases. The carrier cell, which is biased in class-AB, has gain compression at high output power levels, while the class-C biased peak cell has gain expansion. Hence, the gain expansion of the peak cell can compensate for the gain compression of the carrier cell. Specifically, the third-order intermodulation (IM3) level from the carrier cell increases and the phase of IM3 is decreased because the gain of the carrier cell is compressed. On the other hand, when the gain of the peak cell is expanded with uneven drive, both the IM3 level and phase increase. To cancel out the IM3s from the two cells, the components must be 180° out-of-phase with the same amplitudes.



▲ *Fig. 5 Inverted Doherty power amplifier topology.*

Inverted Doherty Topology

The carrier amplifier operational theories indicated that the best efficiency at average envelope power actually occurred with a load impedance closer to 25 than 100 Ω. In order to achieve maximum efficiency at 100 Ω, approximately one-quarter wavelength of a 50 Ω transmission line will be introduced in the carrier amplifier's output matching network (OMN_c). Similarly, the off-state impedance presented by the peak amplifier is so low that this also suggests appending a λ/4 wavelength 50 Ω line to the peak amplifier's output matching network (OMN_p) to guarantee high impedance at the combining node. Size and loss constraints make this approach undesirable. By reversing the Doherty combining point, a 25 Ω maximum efficiency load is provided for the carrier amplifier at the average envelope power. The impedance inversion previously accomplished with the 50 Ω, λ/4 line is incorporated into OMN_c, which constrains $\theta S_{21} = -90°$. In the peak cell, a 50 Ω, λ/4 line becomes the off-state impedance rotation appended to OMN_p. The output is then taken from the carrier amplifier side of the combining node, as shown in ***Figure 5***. This configuration is called an "inverted Doherty" power amplifier.

This "inverted Doherty" will guarantee a high efficiency at low drive level. However, the biggest challenge of the Doherty design is the carrier amplifier output match. In addition to the –90° phase requirement mentioned above, the gain of the carrier amplifier must decrease by 3 dB as its output power transitions between average envelope power and half of the peak envelope power. This can be understood by noting that the carrier amplifier's input power ranges from average envelope power to peak envelope power, while the required output power range is from average envelope power to half of the peak envelope power. The gain reduction is necessary to accommodate half of the peak envelope power of the peaking amplifier. With uneven drive, more power will be delivered to the peak cell. This creates a constant gain for the composite Doherty amplifier at lower power gain, which is an important linearity consideration. To optimize the

WHITEPAPER

Agilent Technologies

De-mystifying Single Carrier FDMA
The New LTE Uplink

Third-generation wireless communication systems based on W-CDMA (wideband code-division multiple access) are being deployed all over the world. To ensure that these systems remain competitive, the 3GPP (3rd Generation Partnership Project) initiated a project in late 2004 for the long-term evolution (LTE) of 3GPP cellular technology.

This article focuses on the physical layer ("Layer 1") characteristics of the LTE uplink, describing the new Single-Carrier Frequency Division Multiple Access (SC-FDMA) transmission scheme and some of the measurements associated with it. Understanding the details of this new transmission scheme and measurements is a vital step towards developing LTE UE designs and getting them to market.

Written by: Moray Rumney BSc, C. Eng, MIET
Lead Technologist, Agilent Technologies

The LTE specifications are being documented in Release 8 of the 3GPP standard. The core specifications are scheduled to be completed by mid-2008 with the conformance test specifications following approximately six months later.

With early system deployment expected in the 2010 timeframe, LTE provides a framework for an evolved 3G network, and aims specifically to achieve the following:

- Increased uplink peak data rates up to 86.4 Mbps in a 20 MHz bandwidth with 64QAM (quadrature amplitude modulation)
- Increased downlink peak data rates up to 172.8 Mbps in a 20 MHz bandwidth with 64QAM and 2x2 SU-MIMO (single-user multiple input/multiple output)
- Maximum downlink peak data rates up to 326.4 Mbps using 4x4 SU-MIMO
- Spectrum flexibility with scalable uplink and downlink channel bandwidths from 1.4 MHz up to 20 MHz
- Improved spectral efficiency, with a 2-4 times improvement over Release 6 HSPA (high speed packet access)
- Sub-5 ms latency for small IP (internet protocol) packets
- Mobility optimized for low mobile speed from 0 to 15 km/h; higher mobile speeds up to 120 km/h will be supported with high performance with the system operating up to 350 km/h
- Co-existence with legacy systems while evolving towards an all-IP network

The LTE Air Interface

There are two primary duplexing modes used in LTE which are frequency division duplex (FDD) and time division duplex (TDD). Variants including half-rate FDD are also anticipated. The integration of the FDD and TDD modes of LTE is much closer than was the case with UMTS. The downlink transmission scheme is based on orthogonal frequency division multiplexing (OFDM) and the uplink uses a new transmission scheme called SC-FDMA. This new scheme borrows from both traditional single-carrier schemes as well as from OFDM.

OFDM and OFDMA

OFDM has been around since the mid 1960s and is now used in a number of non-cellular wireless systems such as Digital Video Broadcast (DVB), Digital Audio Broadcast (DAB), Asymmetric Digital Subscriber Line (ADSL) and some of the 802.11 family of Wi-Fi standards. OFDM's adoption into mobile wireless has been delayed for two main reasons. The first is the sheer processing power which is required to perform the necessary FFT operations. However, the continuing advance of signal processing technology means that this is no longer a reason to avoid OFDM, and it now forms the basis of the LTE downlink. The other reason OFDM has been avoided in mobile systems is the very high peak to average ratio (PAR) signals it creates due to the parallel transmission of many hundreds of closely-spaced subcarriers. For mobile devices this high PAR is problematic for both power amplifier design and battery consumption, and it is this concern which led 3GPP to develop the new SC-FDMA transmission scheme.

Multiple access in the LTE downlink is achieved by using an elaboration of pure OFDM called orthogonal frequency division multiple access (OFDMA). This method allows subcarriers to be allocated to different users. This facilitates the trunking of many lower-rate users as well as enabling the use of frequency hopping to mitigate the effects of narrowband fading.

SC-FDMA

SC-FDMA is a hybrid transmission scheme which combines the low PAR characteristics of single-carrier transmission systems - such as those used for GSM and CDMA - with the long symbol time and flexible frequency allocation of OFDM. The principles behind SC-

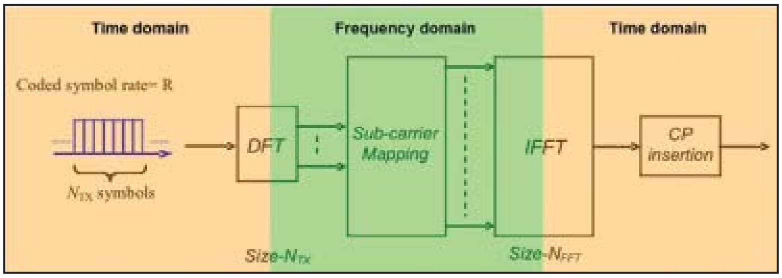


Figure 1 *SC-FDMA signal generation*

FDMA signal generation are shown in Figure 1. This is taken from Figure 1 of the study phase report for the LTE physical layer 3GPP TR 25.814.

On the left hand side of Figure 1 the data symbols are depicted in the time domain. The symbols are converted to the frequency domain using an FFT, and then in the frequency domain they are mapped to the desired location in the overall carrier bandwidth. They must then be converted back to the time domain in order to have the cyclic prefix inserted prior to transmission. An alternative name for SC-FDMA is Discrete Fourier Transform Spread OFDM (DFT-SOFDM).

An alternative description is provided in Figure 2 which shows, in frequency and time, how OFDMA and SC-FDMA would each transmit a sequence of 8 QPSK data symbols. For this simplified example, the number of subcarriers (M) is set to four. For OFDMA, four (M) symbols are taken in parallel, each of them modulating its own subcarrier at the appropriate QPSK phase. Each data symbol occupies 15 kHz for the period of one OFDMA symbol which lasts for 66.7µs. At the start of the next OFDMA symbol, the guard interval containing the cyclic prefix (CP) is inserted. The CP is a copy of the end of a symbol prepended to the start of the symbol. Due to the parallel transmission, the data symbols are the same length as the OFDMA symbols.

In the SC-FDMA case, the data symbols are transmitted sequentially. Since this example involves four subcarriers, four data symbols are transmitted sequentially in one SC-FDMA symbol period. The SC-FDMA symbol period is the same length as the OFDMA symbol at 66.7µs but due to sequential transmission, the data symbols are shorter being 66.7/M µs. A consequence of the higher data rate symbols means more bandwidth is required, so each data symbol occupies 60 kHz of spectrum rather than the 15 kHz for the slower data symbols used for OFDMA. After the four data symbols have been transmitted, the CP is inserted.

Following this graphical comparison of OFDMA and SC-FDMA, the detail of the SC-FDMA signal generation process is shown in Figures 3 and 4. A time domain representation of the data symbol sequence is first generated as shown in Figure 3.

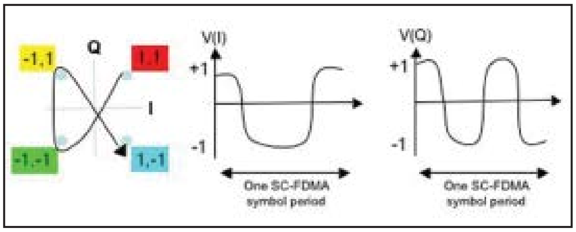


Figure 3 *Creating the time-domain waveform of an SC-FDMA symbol*

For this four subcarrier example a sequence of four data symbols is required to generate one SC-FDMA symbol. Using the first four of the color-coded QPSK data symbols from Figure 2, the process creates one SC-FDMA symbol in the time domain by computing the trajectory traced by moving from one QPSK data symbol to the next. This is done at M times the rate of the SC-FDMA symbol such that one SC-FDMA symbol contains M consecutive QPSK data symbols. For simplicity, we will not discuss time-domain filtering of the data symbol transitions even though such filtering will be present in any real implementation.

Having created an IQ representation in the time domain of one SC-FDMA symbol, the next stage is to represent this in the frequency domain using a discrete Fourier transform (DFT; Figure 4).

The DFT sampling frequency is chosen such that the time-domain waveform of one SC-FDMA symbol is fully represented by M DFT bins spaced 15 kHz apart, where each bin represents one subcarrier with amplitude and phase held

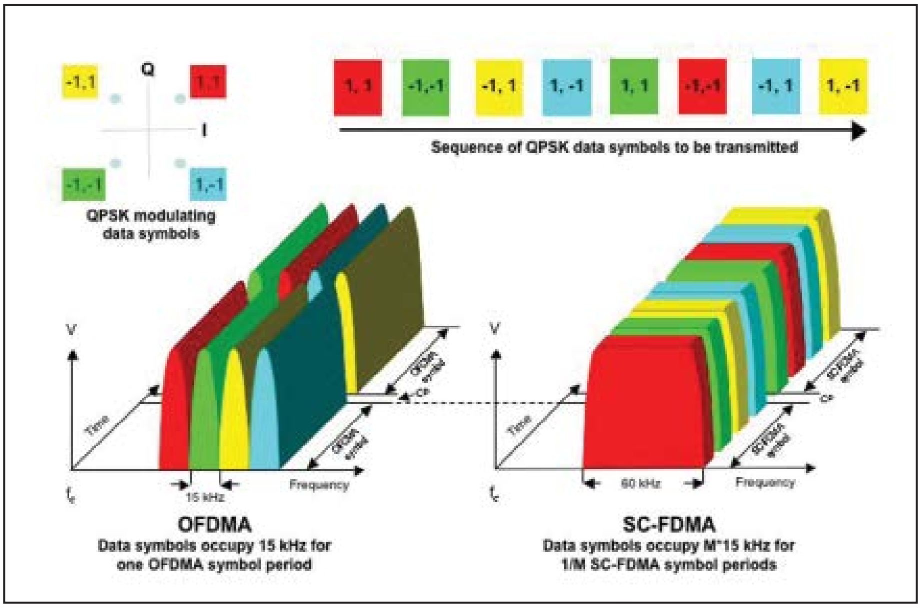


Figure 2 *Comparison of OFDMA and SC-FDMA transmitting a series of QPSK data symbols*

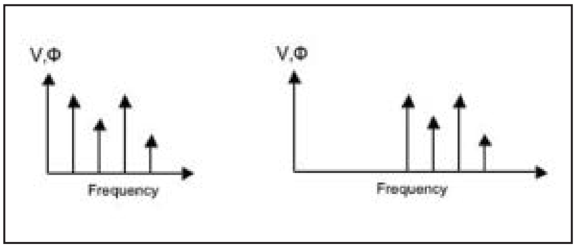


Figure 4. *Baseband and shifted frequency domain representations of an SC-FDMA symbol*

constant for the 66.7μs SC-FDMA symbol period. There is always a one-to-one correlation between the number of data symbols to be transmitted during one SC-FDMA symbol period and the number of DFT bins created — which in turn becomes the number of occupied subcarriers. When an increasing number of data symbols are transmitted during one SC-FDMA period, the time-domain waveform changes faster, generating a higher bandwidth and hence requiring more DFT bins to fully represent the signal in the frequency domain.

Multipath Resistance With Short Data Symbols?

At this point it is reasonable to ask, "How can SC-FDMA still be resistant to multipath when the data symbols are still short?" In OFDMA, the modulating data symbols are constant over the 66.7 μs OFDMA symbol period but an SC-FDMA symbol is not constant over time since it contains M data symbols of much shorter duration. The multipath resistance of the OFDMA demodulation process seems to rely on the long data symbols that map directly onto the subcarriers. Fortunately, it is the constant nature of each subcarrier— not the data symbols — that provides the resistance to delay spread. As shown earlier, the DFT of the time-varying SC-FDMA symbol generated a set of DFT bins constant in time during the SC-FDMA symbol period even though the modulating data symbols varied over the same period. It is inherent to the DFT process that the time-varying SC-FDMA symbol - made of M serial data symbols - is represented in the frequency domain by M time-invariant subcarriers. Thus, even SC-FDMA with its short data symbols benefits from multipath protection. Figure 2 shows the SC-FDMA subcarriers all at the same amplitude but in reality each will have its own amplitude and phase for any one SC-FDMA symbol period.

To conclude SC-FDMA signal generation, the process follows the same steps as for OFDMA. Performing an inverse FFT converts the frequency-shifted signal to the time domain and inserting the CP provides OFDMA's fundamental robustness against multipath.

Figure 5 shows the close relationship between SC-FDMA and OFDMA. The orange blocks represent OFDMA processing and the blue blocks represent the additional time domain processing required for SC-FDMA.

UL Signals	Full Name	Purpose
DMRS	(Demodulation) Reference Signal	Used by the base station for synchronization to the UE and for UL channel estimation. Associated with PUCCH or PUSCH
SRS	Sounding Reference Signal	Used for channel estimation when there is no PUCCH or PUSCH
UL Channels	**Full name**	**Purpose**
PRACH	Physical Randon Access Channel	Call setup
PUCCH	Physical Uplink Control Channel	Scheduling, ACK/NACK
PUSCH	Physical Uplink Shared Channel	Payload

Table 1 *Uplink signals and channels*

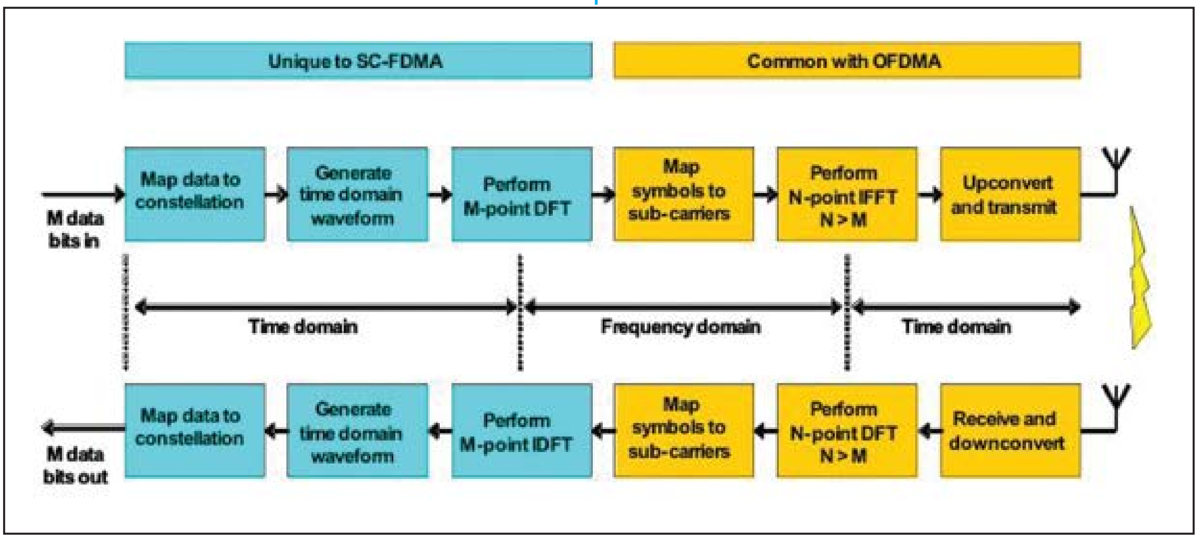


Figure 5 *Simplified model of SC-FDMA generation and reception*

The key point to note is that the signal which is converted from the frequency domain back to the time domain is no more than a frequency shifted version of a series of QPSK symbols. This example illustrates the main reason SC-FDMA was developed: that is, the PAR of the final signal is no worse than that of the original data symbols, which in this case was QPSK. This is very different to OFDMA where the parallel transmission of the same QPSK data symbols creates statistical peaks - much like Gaussian noise - far in excess of the PAR of the data symbols themselves. Limiting PAR using SC-FDMA significantly reduces the need for the mobile device to handle high peak power. This lowers costs and reduces battery drain.

Physical Layer Structure

The LTE physical layer comprises two types of signals known as physical signals and physical channels. Physical signals are generated in Layer 1 and used for system synchronization, cell identification, and radio channel estimation. Physical channels carry data from higher layers including control, scheduling, and user payload. Table 1 shows the uplink physical signals and channels.

Uplink Frame Structure

There are two uplink frame structures, one for FDD operation called type 1 and the other for FDD operation called type 2. Frame structure type 1 is 10 ms long and consists of ten subframes, each

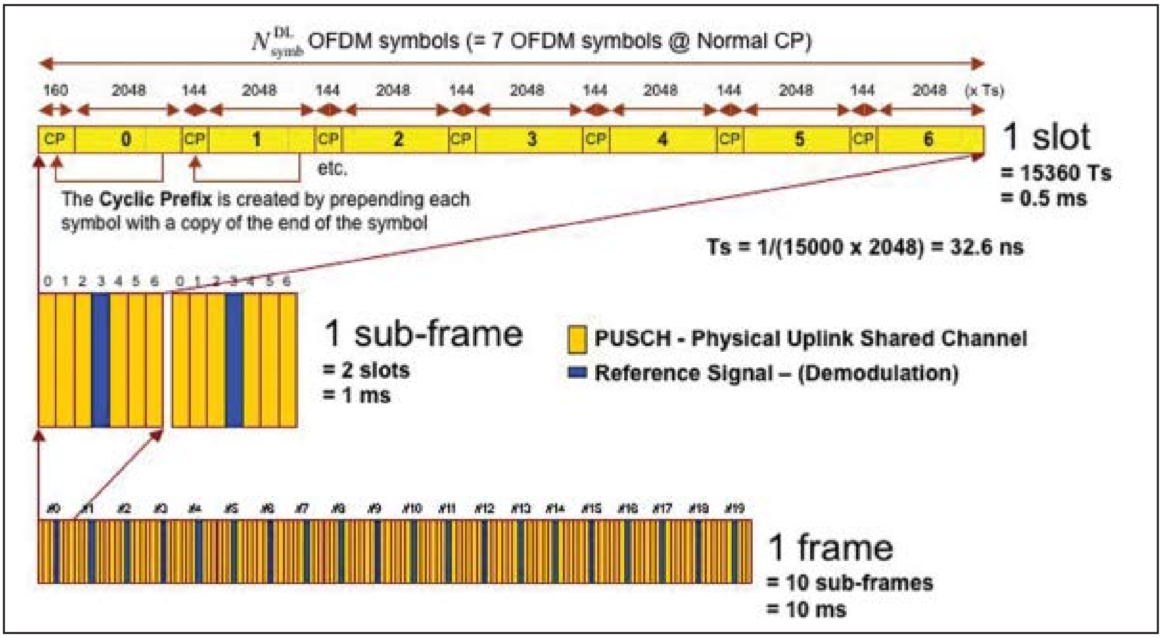


Figure 6 *Frame Structure 1 for uplink showing mapping for DMRS and PUSCH*

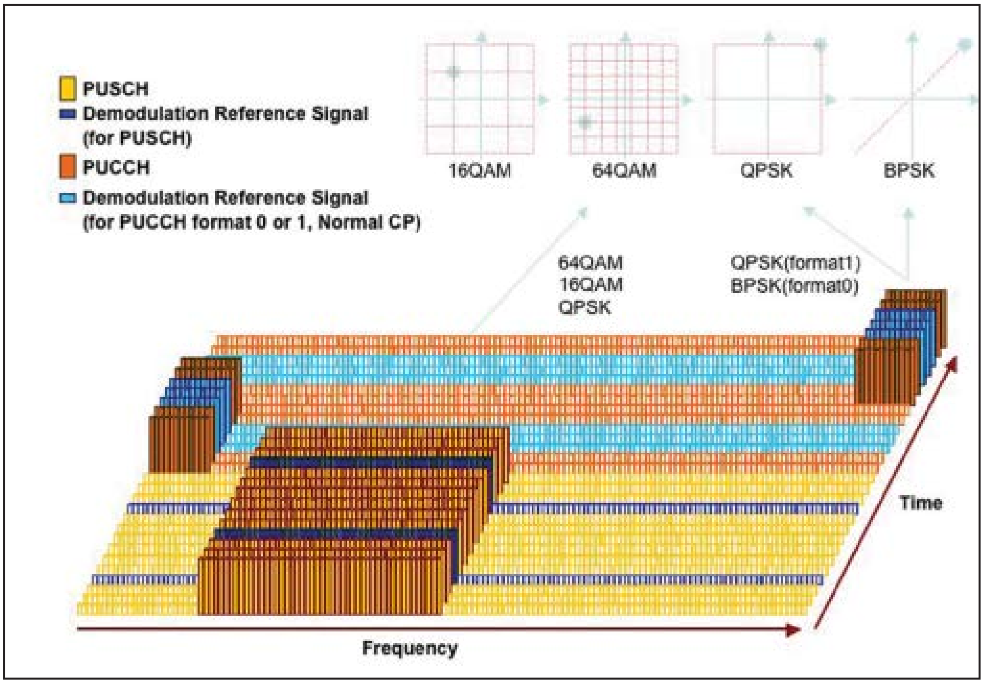


Figure 7 *Frame Structure 1 for the uplink showing one subframe vs. frequency*

***Figure 8** Analysis of a 16QAM SC-FDMA signal*

comprising two 0.5 ms slots. Figure 6 shows how the DMRS and PUSCH map onto the frame structure. The number of symbols in a slot depends on the CP length. For a normal CP, there are seven SC-FDMA symbols per slot. For an extended CP used for when the delay spread is large, there are six SC-FDMA symbols per slot.

Demodulation reference signals are transmitted in the fourth symbol (that is, symbol number 3) of every slot. The PUSCH can be transmitted in any other symbol.

Figure 7 shows the uplink frame structure type 1 in both frequency and time. Each vertical bar represents one subcarrier. Transmissions are allocated in units called resource blocks (RB) comprising 12 adjacent subcarriers for a period of 0.5 ms. In addition to the DMRS and PUSCH the figure also shows the PUCCH which is always allocated to the edge RB of the channel bandwidth alternating from low to high frequency on adjacent slots. Note that the frequency allocation for one UE is typically less than the system bandwidth. This is because the number of RB allocated directly scales to the transmitted data rate which may not always be the maximum. The DMRS is only transmitted within the PUSCH and PUCCH frequency allocation-unlike the reference signals on the downlink which are always transmitted across the entire channel bandwidth even if the channel is not fully occupied.

If the base station needs to estimate the uplink channel conditions when no control or payload data is scheduled then it will allocate the SRS which is independent of the PUSCH and PUCCH. The PUSCH can be modulated at QPSK, 16QAM or 64QAM. The PUCCH is only QPSK and the DMRS is BPSK with a 45 degree rotation.

Analyzing an SC-FDMA Signal

Figure 8 shows some of the measurements that can be made on a typical SC-FDMA signal using the Agilent 89601A Vector Signal Analyzer software. The IQ constellation in trace A (top left) shows that this is a 16QAM signal. The unity circle represents the DMRS occurring every seventh symbol, which are phase-modulated using an orthogonal Zadoff-Chu sequence.

Trace B (lower left) shows signal power versus frequency. The frequency scale is in 15 kHz sub-carriers numbered from -600 to 599, which represents a bandwidth of 18 MHz or 100 RB. The nominal channel bandwidth is therefore 20 MHz and the allocated signal bandwidth is 5 MHz towards the lower end. The brown dots represent the instantaneous subcarrier amplitude and the white dots the average over 10 ms. In the center of the trace, the spike represents the local oscillator (LO) leakage - IQ offset - of the signal; the large image to the right is an OFDM artifact deliberately created using 0.5 dB IQ gain imbalance in the signal. Both the LO leakage and the power in non-allocated sub-carriers will be limited by the 3GPP specifications.

Trace C (top middle) shows a summary of the measured impairments including the error vector magnitude (EVM), frequency error, and IQ offset. Note the data EVM at 1.15 percent is much higher than the DMRS EVM at 0.114 percent. This is due to a +0.1 dB boost in the data power as reported in trace E, which for this example was ignored by the receiver to create data-specific EVM. Also note the DMRS power boost is reported as +1 dB, which can also be observed in the IQ constellation because the unity circle does not pass through eight of the 16QAM points. Trace D (lower middle) shows the distribution of EVM by subcarrier. The average and peak of the allocated signal EVM is in line with the numbers in trace C. The EVM for the non-allocated subcarriers reads much higher, although this impairment will be specified with a new “in-band emission” requirement as a power ratio between the allocated RB and unallocated RB. The ratio for this particular signal is around 30 dB as trace B shows. The blue dots in trace D also show the EVM of the DMRS, which is very low.

Figure 9 *Agilent's MXG Vector Signal Generator with LTE Signal Studio software and the MXA Signal Analyzer with 89600 LTE VSA software provides the most comprehensive solution for physical layer testing.*

Trace E (top right) shows a measurement of EVM by modulation type from one capture. This signal uses only the DMRS phase modulation and 16QAM so the QPSK and 64QAM results are blank. Finally, trace F (lower right) shows the PAR — the whole point of SC-FDMA — in the form of a complementary cumulative distribution function (CCDF) measurement. It is not possible to come up with a single figure of merit for the PAR advantage of SC-FDMA over OFDMA because it depends on the data rate. The PAR of OFDMA is always higher than SC-FDMA even for narrow frequency allocations; however, when data rates rise and the frequency allocation gets wider, the SC-FDMA PAR remains constant but OFDMA gets worse and approaches Gaussian noise. A 5 MHz OFDMA 16QAM signal would look very much like Gaussian noise. From the white trace it can be seen at 0.01 percent probability the SC-FDMA signal is 3 dB better than the blue Gaussian reference trace. As every amplifier designer knows, shaving even a tenth of a decibel shaved from the peak power budget is a significant improvement.

Agilent Design and Test Solutions to Help You Take LTE Forward

As a world leader in test and measurement solutions, Agilent Technologies is at the forefront of emerging wireless and broadband markets, such as LTE. Agilent provides the most complete suite of LTE tools, offering design and test solutions for the entire product development cycle – from RF and digital early design & test through protocol development to network deployment.

Included in this comprehensive suite of LTE tools are solutions to design and simulate LTE signals, create and measure LTE encoded signals with sources and analyzers, and test mixed analog & digital signals – see figure 9. Just added to Agilent's suite of LTE solutions is a one-box tester that provides the platform for protocol design and test solutions, in partnership with Anite. This platform will provide RF and protocol conformance test systems when they are needed. And, the newly introduced signaling analyzer enables analysis of the new LTE/SAE network.

So as you take LTE forward, Agilent will continue to clear the way.

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Agilent Measurement Journal, Issue 4 2008

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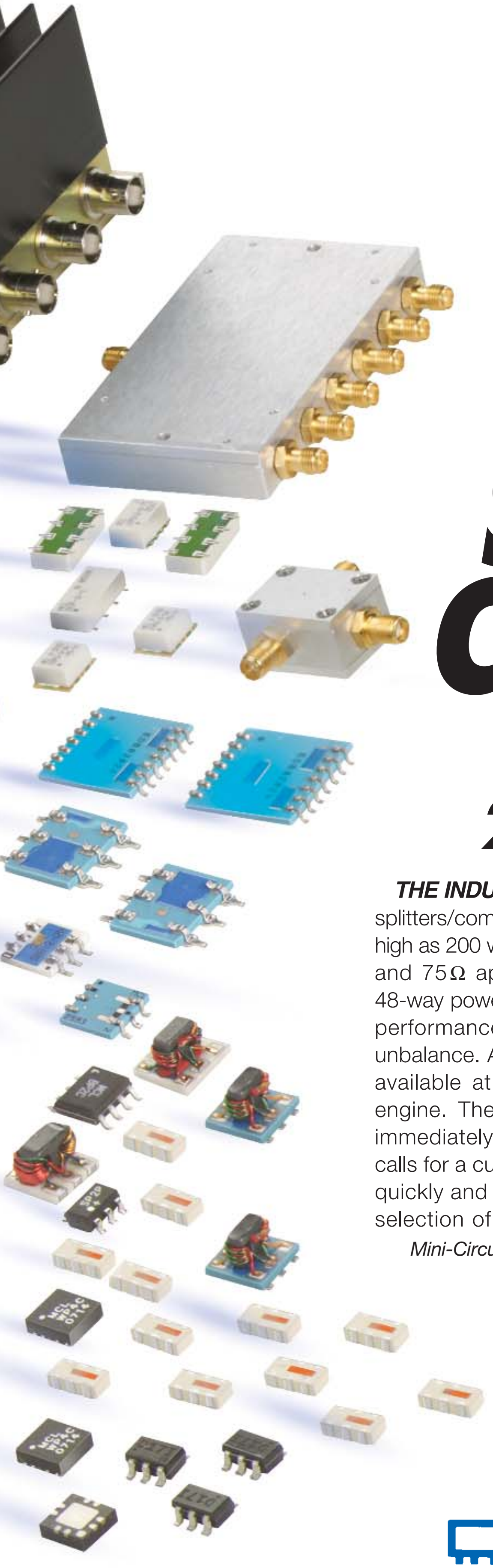
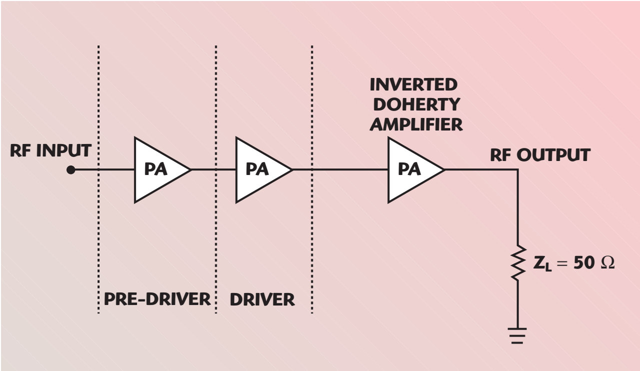


Doherty amplifier's average efficiency, the carrier amplifier's output match must be designed for best efficiency performance at average envelope power. For the inverted topology, this occurs at 25 Ω. The maximum carrier amplifier efficiency is limited by the linearity as a result of its operation, together with the class-C peaking amplifier. The carrier amplifier design is thus constrained by the gain, phase, efficiency, linearity and absolute power requirements.

Operating in class-C, its transfer characteristic must be smooth, without evidence of discontinuities. The adjacent channel leakage and IMD problems are made obvious with a two-tone test. The design of the bypass and decoupling networks, as well as the bias circuit, are crucial to avoid the problem of bypass capacitors. The output contribution of the peaking amplifier is expected to range from zero to half of peak envelope power for the same drive range, which causes the carrier amplifier to deliver average envelope power to half of peak envelope power. Finally, the phase matching network, Φ is 83.88°, or an electrical length of 0.223λ, determined initially from individual measurements of the carrier and peaking amplifier, when delivering half of the peak envelope power. It is set to provide equal phase lengths in both signal paths. The final phase length is optimized for best linearity and gain flatness.

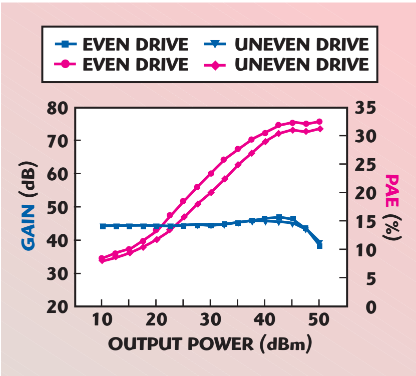
THREE-STAGE STRUCTURE

In this article, a Doherty power amplifier with three stages is proposed because the conventional Doherty power amplifier always has a single stage and its gain is not high enough, from 8 to 15 dB. ***Figure 6*** is a schematic of the proposed Doherty power amplifier. The novel Doherty amplifier consists of three stages: pre-driver, driver and final stage. The pre-driver and driver are used to obtain a higher output and power gain. They both work in class-A, since this method makes the two stages work in the small-signal regime. This can maintain a low noise figure for the entire amplifier, while it also makes the total PAE a little less. The PAE of the entire amplifier depends mainly on the efficiency of the final stage.



▲ *Fig. 6 Three-stage power amplifier.*

▲ *Fig. 7 The Doherty amplifier.*



▲ *Fig. 8 Gain and power-added efficiency vs. output power.*

AMPLIFIER IMPLEMENTATION AND RESULTS

A photograph of the power amplifier is shown in ***Figure 7***. The Doherty power amplifier performance not only considers the linearity, but also its efficiency. To satisfy these demands, a Freescale MFR6S21050L device was used. A conventional Doherty amplifier was also fabricated for comparison.

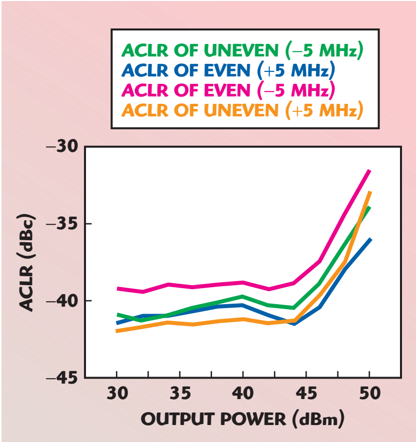
Figure 8 shows the gain and PAE of the proposed Doherty power amplifier and of an ordinary Doherty power amplifier with even drive. As with an uneven power drive (1:2.5), the carrier cell is compressed early and the peaking cell is expanded early and the region is wider than for the usual Doherty amplifier. Therefore, the amplifier with an uneven drive generates a more linear power because the early gain expansion of the peaking cell compensates for the gain compression of the carrier cell over a wide power range. The power gain of the uneven case keeps a better linearity compared to the even drive case. The PAE of the uneven drive amplifier achieves 33 percent, a 3.2 percent improvement from the inverted structure with optimum load impedance modulation.

Figure 9 shows the measured ACLR of the two types of Doherty amplifier. Compared with the even case, the uneven case, with an inverted output structure, delivers a improved ACLR performance by 2 dB and 3.1 dB at the average output power of 45 dBm for a two-tone test at –5 and +5 MHz tone spacing, respectively. These results confirm that the proposed bypass and decoupling networks in OMN_P provides a better adjacent channel leakage performance.

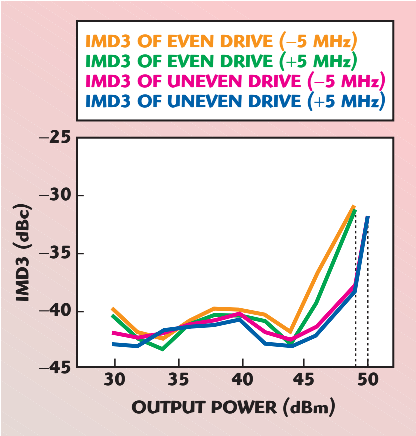
Figure 10 shows the measured IMD3 of the Doherty amplifier with both even and uneven power drives for a two-tone signal. A two-tone signal is used with –5 and +5 MHz tone spacing, respectively. The IMD3 of the amplifier with uneven drive is improved by 10 W compared to the even case.

CONCLUSION

A high linearity and efficiency Doherty amplifier is proposed and fabricated with uneven power drive and inverted topology. Its PAE achieves 33 percent, an improvement



▲ *Fig. 9 ACLR performance for two-tone signals spaced ±5 MHz.*



▲ *Fig. 10 IMD3 for two-tone signal with ±5 MHz spacing.*

of 3.2 percent over the even drive case. With its inverted structure and offset line in the output-matching network, the amplifier offers a better linearity and efficiency than for the even case. The uneven power drive delivers more linear power, which is improved by a better load impedance modulation. For a ±5 MHz two-tone test, the ACLR is –42 dBc at 40 dBm output power, which is a 2 dB improvement. The IMD3 of the proposed amplifier gives a 10 W improvement over the even case, with appropriate cancellation of the carrier cell harmonics. These experimental results clearly demonstrate the superior performance of the proposed Doherty power amplifier, compared to the conventional Doherty power amplifier. The proposed design methods are suited for the design of the Doherty amplifier for high efficiency and high linearity operation. ■

ACKNOWLEDGMENT

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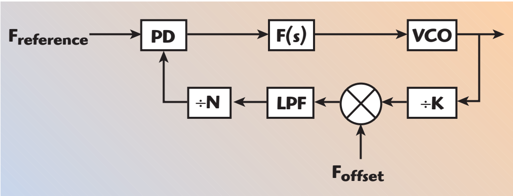
A SELF-OFFSET PHASE-LOCKED LOOP

The quest for a low-noise, high-switching-speed synthesizer has become a common trend in modern wireless systems. A large division ratio is required for the feedback path of a conventional phase-locked loop (PLL) if microwave frequencies need to be synthesized. To mitigate the problem, an offset PLL architecture is often used, which includes a mixer in the feedback path. The output signal from a voltage-controlled oscillator (VCO) is down-converted to a much lower frequency by mixing it with an externally generated offset frequency signal. A wide tuning range, often required in microwave bands, dictates the use of a pseudo-coherent multi-VCO configuration.[1] A single-VCO 10 GHz offset PLL architecture, intended for point-to-point radio applications, has also been reported.[2] On the other hand, a relatively high reference frequency is usually required to obtain similar performance using a conventional PLL.[3]

Fig. 1 Typical microwave frequency synthesizer architecture using an offset PLL.



MICROWAVE FREQUENCY SYNTHESIS

A simplified offset PLL configuration for microwave frequency synthesis is shown in ***Figure 1***. The VCO output signal F_{output} is first divided by a fixed ratio prescaler, having a division ratio K (where K is an integer). The output signal from the prescaler drives an input port of the mixer. An externally generated signal F_{offset} drives the other input (LO port) of the mixer. The resulting low-frequency mixing product is filtered by a low pass filter (LPF) and applied to the main divider, having a division ratio N (where N is generally a variable of real or integer type). A phase detector (PD) compares phase of the feedback signal (from the output of the main divider) with the phase of an externally generated reference signal $F_{reference}$. The resulting error signal is applied, through the loop filter F(s), to the control input of the VCO. Optionally, depending

BOGDAN SADOWSKI
Harris Stratex Networks Inc.
Morrisville, NC

on the technology used, the mixer could be driven directly from the VCO output (in such case the fixed prescaler is omitted).

SELF-OFFSET PLL

The self-offset PLL configuration, proposed here, is shown in ***Figure 2***. Two input signals for the mixer are generated internally within the same PLL, thus the separate offset frequency source is not required. Two independent prescalers (having division ratios K and L, respectively) divide the VCO output frequency to generate the two input signals required for down-conversion mixing. Preferably, the LO port of the mixer is being driven by the prescaler with the larger division ratio L. The difference between the division ratios of the two fixed prescalers is defined as an integer value x such that

$$L = K + x \tag{1}$$

The frequency of the down-converted signal (at the output of the low pass filter), can be expressed by

$$F_{LPF} = \left(\frac{F_{VCO}}{K}\right) - \left(\frac{F_{VCO}}{L}\right) \tag{2}$$

Substituting L from Equation 1 into Equation 2 yields

$$F_{LPF} = \left(\frac{F_{VCO}}{K}\right) - \left[\frac{F_{VCO}}{(K+x)}\right] \tag{3}$$

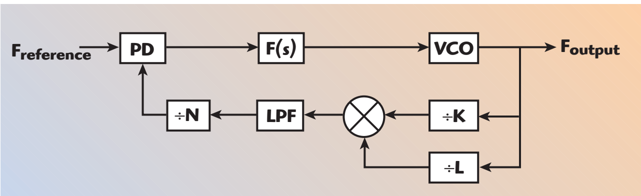
$$F_{LPF} = F_{VCO}\left[\frac{x}{K(K+x)}\right] \tag{4}$$

Therefore, the tuning range "seen" by the main divider is reduced by the factor of

$$R = \frac{x}{K(K+x)} \tag{5}$$

The frequency of the signal at the phase detector's feedback input (at the output of the main divider) can be expressed by

$$F_{PFD} = \frac{F_{LPF}}{N} \tag{6}$$



▲ *Fig. 2 Self-offset PLL.*

Substituting Equation 4 into Equation 6 yields

$$F_{PFD} = F_{VCO}\left[\frac{x}{K(K+x)}\right]\left(\frac{1}{N}\right) \tag{7}$$

For the case of a conventional phase-locked loop, having only a divide by K prescaler and a main divider in the feedback path, the frequency of the signal at the phase detector's feedback input is given as

$$F_{PFD} = \frac{F_{VCO}}{KN_{conv}} \tag{8}$$

where N_{conv} represents the division ratio for the main divider of the conventional PLL.

Assuming identical comparison frequencies for the self-offset and conventional phase-locked loops yields

$$\left(\frac{F_{VCO}}{N}\right)\left[\frac{x}{K(K+x)}\right] = \frac{F_{VCO}}{KN_{conv}} \tag{9}$$

and

$$\frac{\left(\dfrac{F_{VCO}}{KN_{conv}}\right)}{\left(\dfrac{F_{VCO}}{N}\right)} = \left[\frac{x}{K(K+x)}\right] \tag{10}$$

The reduction in the division ratio of the main divider (when comparing the self-offset PLL with the conventional phase-locked loop) can be calculated by further simplifying Equation 10 such that

$$\frac{N}{KN_{conv}} = \left[\frac{x}{K(K+x)}\right] \tag{11}$$

and

$$\frac{N}{N_{conv}} = \frac{x}{(K+x)} \tag{12}$$

The phase noise improvement inside the PLL loop bandwidth, resulting from that reduction, can be estimated by 20 log (N/N_{conv}). ***Figure 3*** shows the phase noise improvement (in dB) as a function of x, with K as a parameter.

The choice of a low value of x and a large value of K gives the largest improvement of the phase noise in dB. Increasing the value of K beyond 10 results in smaller incremental improvements than when varying the value of K from 1 to 10.

As in any conversion scheme involving a mixer, spurious mixing products have to be considered and mitigated. This restricts the choice of the x and K values combination. Mixing products that fall exactly into the frequency band of the desired output are difficult to mitigate. It is therefore important to avoid values of x and K which would result in their generation, as can be described by Equations 13a or b:

$$n\left(\frac{F_{VCO}}{K}\right) - m\left[\frac{F_{VCO}}{(K+x)}\right] \neq F_{LPF} \tag{13a}$$

$$n\left(\frac{F_{VCO}}{K}\right) + m\left[\frac{F_{VCO}}{(K+x)}\right] \neq F_{LPF} \tag{13b}$$

where m, n are the harmonic indexes

Substituting Equation 4 into Equations 13a and b yields

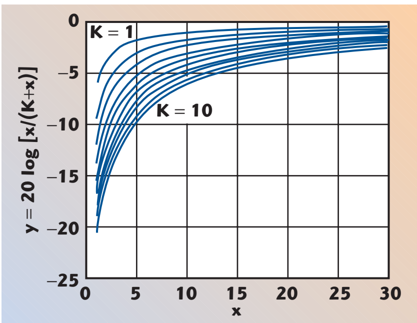
$$n\left(\frac{F_{VCO}}{K}\right) - m\left[\frac{F_{VCO}}{(K+x)}\right] \neq F_{VCO}\left[\frac{x}{K(K+x)}\right] \tag{14a}$$

$$n\left(\frac{F_{VCO}}{K}\right) + m\left[\frac{F_{VCO}}{(K+x)}\right] \neq F_{VCO}\left[\frac{x}{K(K+x)}\right] \tag{14b}$$

From Equations 14a and b

$$\left(\frac{n}{K}\right) - \left(\frac{m}{K+x}\right) \neq \frac{x}{K(K+x)} \tag{15a}$$

$$\left(\frac{n}{K}\right) + \left(\frac{m}{K+x}\right) \neq \frac{x}{K(K+x)} \tag{15b}$$



▲ *Fig. 3 Phase noise improvement due to reduction in the main divider's division ratio.*

Then,

$$\frac{n(K+x)-mK}{K(K+x)} \neq \frac{x}{K(K+x)} \quad (16a)$$

$$\frac{n(K+x)+mK}{K(K+x)} \neq \frac{x}{K(K+x)} \quad (16b)$$

which gives

$$n(K+x)-mK \neq x \quad (17a)$$

$$n(K+x)+mK \neq x \quad (17b)$$

$$(n-m)K \neq (1-n)x \quad (18a)$$

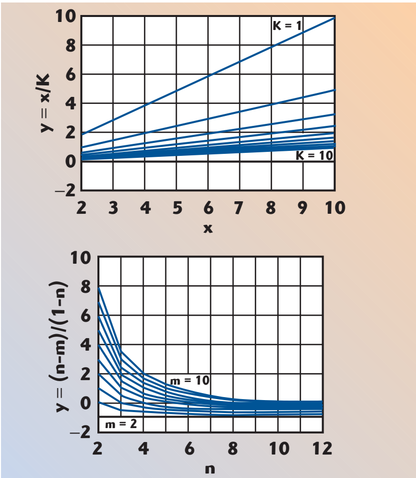
$$(n+m)K \neq (1-n)x \quad (18b)$$

thus

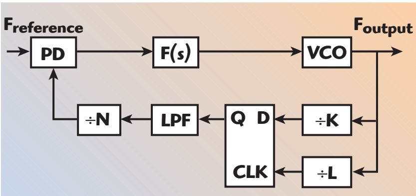
$$\frac{(n-m)}{1-n} \neq \frac{x}{K} \quad (19a)$$

$$\frac{(n+m)}{1-n} \neq \frac{x}{K} \quad (19b)$$

In most cases the x value will be chosen to be much smaller than the K value. Therefore, Equation 19a will be considered here as a design guide and is graphically represented in ***Figure 4***. The right- and left-hand sides of Equation 19a are displayed side by side (the y axis has exactly the same scale on both graphs). Combinations of x, K, which give the same y axis value as any combination of harmonic indexes m, n, are to be avoided. For practical circuits it is not always possible to avoid all spurious products; therefore, some compromise solution has to be pursued. When the division ratio K is made to be a large integer number (in order to maximize the phase noise improvement), mixing can be done using "all digital" implementation, as shown in ***Figure 5***. A D-type flip-flop can be used to produce the mixing product of two signals, driving its D and CLK (clock) inputs. An analog mixer can therefore be replaced with a "digital mixer".[4]



▲ *Fig. 4 Guide for in-band spurious products avoidance.*



▲ *Fig. 5 Self-offset PLL with "digital mixer."*

From Equation 1

$$F_{CLK} < F_D \quad (20)$$

The frequency of the mixing product, delivered by the Q output, can be expressed either as

$$F_Q = F_D - MF_{CLK} \quad (21)$$

where

F_Q = frequency of the signal at Q output
F_{CLK} = frequency of the signal driving CLK input
M = 1,2,3,…
FD = frequency of the signal driving D input on condition that FD is constrained as

$$MF_{CLK} < F_D < (MF_{CLK}) + \left(\frac{F_{CLK}}{2}\right) \quad (22)$$

or expressed as

$$F_Q = [(M+1)F_{CLK}] - F_D \quad (23)$$

on condition that F_D is constrained as

$$(MF_{CLK}) + \left(\frac{F_{CLK}}{2}\right) < F_D < (M+1)F_{CLK} \quad (24)$$

Thus, there are two modes of digital mixer operation: a sideband non-inverting mixing mode and a sideband inverting mixing mode. In order to maintain only one mode of mixing, the frequency of the signal driving the D input has to be constrained. As a result, the choice of K and x values combination is further restricted. As an example, the case of non-inverting

mixing will be analyzed here. Equation 21 can be written as

$$F_Q = \left(\frac{F_{VCO}}{K}\right) - M\left(\frac{F_{VCO}}{L}\right) \qquad (25)$$

$$F_Q = \left(\frac{F_{VCO}}{K}\right) - M\left[\left(\frac{F_{VCO}}{(K+x)}\right)\right] \qquad (26)$$

Therefore, the tuning range seen by the main divider is, in this case, reduced by the factor of

$$R_{digital} = \frac{(K+x) - MK}{K(K+x)} \qquad (27)$$

Thus

$$R_{digital} = \frac{x + K(1-M)}{K(K+x)} \qquad (28)$$

The reduction in the division ratio of the main divider (when comparing a self-offset PLL including a digital mixer with a conventional phase-locked loop) can now be derived as well. Assuming identical comparison frequencies for the digital self-offset and conventional phase-locked loops yields

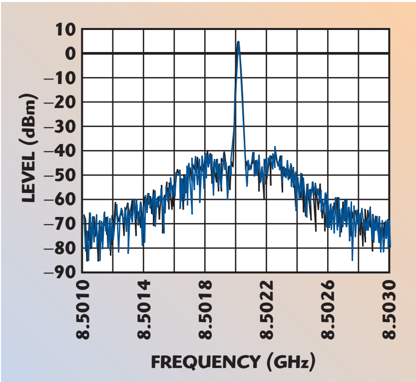
$$\left(\frac{F_{VCO}}{N}\right)\left[\frac{x + K(1-M)}{K(K+x)}\right] = \frac{F_{VCO}}{KN_{conv}} \qquad (29)$$

$$\frac{\left(\frac{F_{VCO}}{KN_{conv}}\right)}{\left(\frac{F_{VCO}}{N}\right)} = \frac{x + K(1-M)}{K(K+x)} \qquad (30)$$

$$\frac{N}{KN_{conv}} = \frac{x + K(1-M)}{K(K+x)} \qquad (31)$$

Then

$$\frac{N}{N_{conv}} = \frac{x + K(1-M)}{K+x} \qquad (32)$$



▲ *Fig. 6 Measured close-in spectrum of the prototype self-offset PLL.*

It is easy to see that Equation 28 is equivalent to Equation 5 and Equation 32 is equivalent to Equation 12 when M = 1. Furthermore, Equation 22 can be written as

$$M\left[\frac{F_{VCO}}{(K+x)}\right] < \left(\frac{F_{VCO}}{K}\right) < \left\{M\left[\frac{F_{VCO}}{K+x}\right]\right\} + \left[\frac{F_{VCO}}{2(K+x)}\right] \qquad (33)$$

$$M\left[\frac{F_{VCO}}{(K+x)}\right] < \left(\frac{F_{VCO}}{K}\right) < \left\{\left[M + \left(\frac{1}{2}\right)\right]\left[\frac{F_{VCO}}{(K+x)}\right]\right\} \qquad (34)$$

which gives

$$M < \left(\frac{K+x}{K}\right) < \left[M + \left(\frac{1}{2}\right)\right] \qquad (35)$$

$$M < \left[1 + \left(\frac{x}{K}\right)\right] < \left[M + \left(\frac{1}{2}\right)\right] \qquad (36)$$

from which it can be concluded that

$$(M-1) < \left(\frac{x}{K}\right) < \left[M - \left(\frac{1}{2}\right)\right] \qquad (37)$$

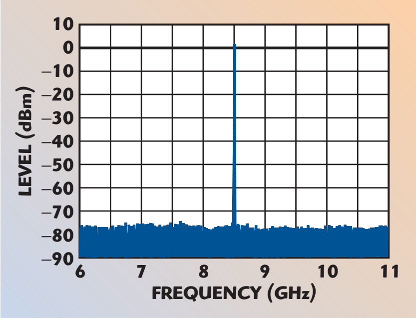
Thus, in order to maintain only one mode (of non-inverting mixing), the choice of K and x values has to be restricted, as described by Equation 37.

EXPERIMENTAL RESULTS

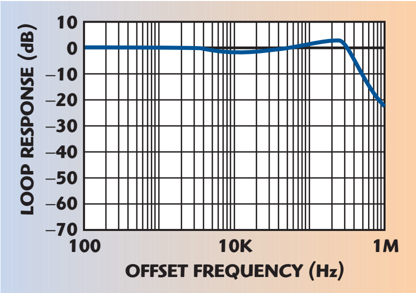
The concept of a self-offset PLL has been experimentally verified for the case that utilizes an analog mixer. The prototype included an HMC510 MMIC VCO from Hittite Microwave Corp. (9 GHz push-push type HBT VCO), which has two additional outputs: a half-frequency and divide-by-four. The half-frequency output was further divided by a divide-by-three prescaler HMC437 (from Hittite Microwave Corp.), while the divide-by-four output was further divided by a divide-by-two prescaler HMC364 (also from Hittite Microwave Corp.). Thus, the combination of K = 6 and x = 2 was used for this prototype arrangement. In order to synthesize the output frequencies from 8.151 to 8.851 GHz, with the comparison frequency set to 100 MHz, division ratios for the main divider shall cover a range between 3.39625 and 3.6879167. To obtain such small divi-

sion ratios, an AD9858 direct digital synthesizer (from Analog Devices Inc.) was used as a fractional-N main divider. The AD9858 DDS was particularly convenient to use, due to its fine frequency resolution and its level of integration. The AD9858 includes an analog (Gilbert cell type) mixer and a 150 MHz phase-frequency detector. The drawback is that the noise floors of the DDS and phase-frequency detector limit the achievable close-in noise pedestal at the PLL output. Other drawbacks are the spurious signals at the DDS output (inherent in DDS). Therefore, a lower than maximum achievable PLL closed-loop bandwidth has to be set in order to mitigate the problem.

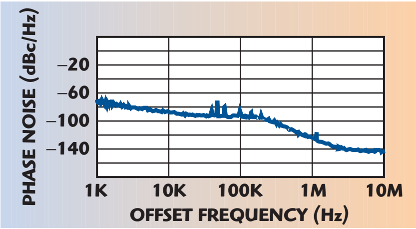
The measured output spectrum of the prototype self-offset PLL (in the vicinity of 8.502 GHz) is shown in ***Figure 6***. Close-in spurious are at least –44 dBc (below carrier) and, further away, at least –77 dBc, as shown in ***Figure 7***. ***Figure 8*** shows the measured closed-loop response. The loop bandwidth was set to 350 kHz. The phase noise of the prototype self-offset PLL (in the vicinity of the top of the tuning range) is shown in ***Figure 9***.



▲ *Fig. 7 Measured output spectrum of the prototype self-offset PLL.*



▲ *Fig. 8 Measured closed-loop response of the prototype self-offset PLL.*



▲ *Fig. 9 Measured phase noise of the prototype self-offset PLL at 8.842 GHz.*

CONCLUSION

A new class of phase-locked loop, the self-offset PLL, has been proposed and experimentally verified. The new configuration permits a large reduction in the division ratio of the main divider (when comparing the self-offset PLL with a conventional phase-locked loop), resulting in an increased open-loop gain and the possibility for fast settling and significant improvement of the phase noise inside the PLL loop bandwith. Two variants (utilizing either analog mixing or digital mixing) were discussed. From a theoretical study, simple design guidelines were derived for each of the two variants. ■

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Bogdan Sadowski *received his MS degree from the Warsaw University of Technology (formerly Warsaw Technical University), Warsaw, Poland, in 1985. He is currently a senior RF engineer with Harris Stratex Networks Inc. (formerly the Microwave Communications Division of Harris Corp.). His interests include frequency synthesis and nonlinear circuits.*

A BROADBAND CLASS-E PARALLEL-CIRCUIT VHF POWER AMPLIFIER WITH HIGH HARMONIC SUPPRESSION

Class-E power amplifiers (PA) offer high efficiency. However, with broadband load networks, their harmonic performance degrades. This article introduces a practical implementation to achieve high harmonic suppression in a broadband class-E amplifier design by combining a parallel-circuit load network with a reactance compensation technique. A three-stage power amplifier is used to provide high gain and output power as required for two-way radio applications. The final PA stage uses a broadband class-E design, which combines a parallel-circuit load network with a reactance compensation technique. A novel series L_xC_x, with a high-Q roll-off harmonic filter (elliptical type), provides high harmonic suppression while maintaining the class-E switching mode over a broad frequency band. An efficiency of 70 percent and a second-harmonic suppression of –73 dBc across a wide bandwidth (135 to 175 MHz), at an operating power of 6.5 W, are demonstrated. Up to now, these harmonic and efficiency results are the highest obtained over a broad VHF frequency band, with a low supply voltage of 7.2 V.

Modern portable two-way radios are required to operate over a large number of channels for a long period of time from a small size battery. This means that an effective battery life solution is necessary. In radio transmitters, the PA is generally the main consumer of battery dc power. Therefore, the best way of improving battery life is to increase the PA efficiency.

Class-E PA circuits are suitable for high efficiency amplification at radio frequency and microwave ranges. The concept of class-E with shunt capacitance was introduced by Sokal.[1] Grebennikov proposed a broadband class-E design,[2] combining a parallel-circuit load network with reactance compensation, which is very suitable for practical two-way radio applications. The use of a low dc-feed inductance, a higher load resistance, a larger shunt capacitance and a higher maximum operating frequency topology are attractive solutions that have some advantages over class-E with only a shunt capacitance.[2,3] However, due to the inherent asymmetrical input drive of class-E, significant harmonic content is generated in the output voltage and current.[4,7]

The conventional design of a high-efficiency switched-mode tuned PA requires a high-

KUMAR NARENDRA,
SANGARAN PRAGASH AND
CHACKO PRAKASH
Motorola Technology
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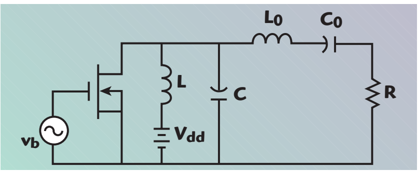
Q_L factor to satisfy the necessary harmonic impedance conditions at the output device terminal.[5] In this article, a series L_xC_x, with an elliptical harmonic filter, provides high harmonic suppression while maintaining the broadband switching mode of class-E. A high harmonic suppression (more than –73 dBc) and efficiency (70 percent) are demonstrated experimentally over a broad frequency range, from 135 to 175 MHz. In a comparable work,[6] Mury demonstrated a second-harmonic suppression of 47 dBc for a power of 22 dBm.

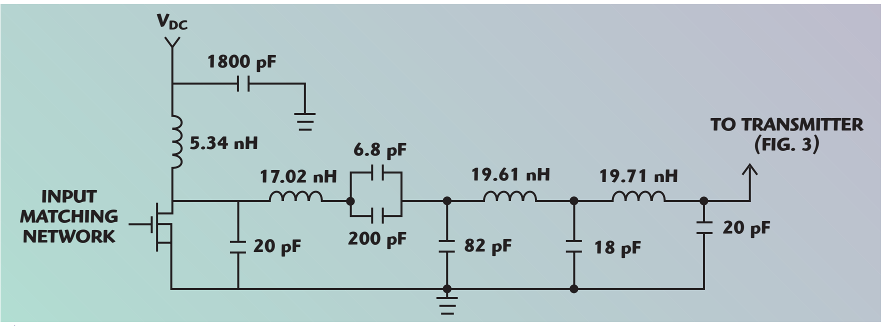
BROADBAND CLASS-E PA WITH SERIES LC AND HIGH-Q ROLL-OFF HARMONIC FILTER FOR HIGH HARMONIC SUPPRESSION

A three-stage amplifier chain was chosen to achieve high gain and high output power. The first stage (ADA4743-HBT device)[9] and second stage (RD01MUS1-LDMOS device)[10] operate in class-AB. A direct matching technique was used for the output match of the second stage and input match of the final stage. The optimum load impedance of the second stage and optimum source impedance of the final stage PA were obtained from simulation and correlated with load-pull measurements. A 7.2 V supply voltage was used for the second and third stages and 5 V was applied to the first stage. The RD07MVS1-LDMOS device[11] was used for the broadband class-E power stage design that combines the parallel-circuit load network with a reactance compensation technique.

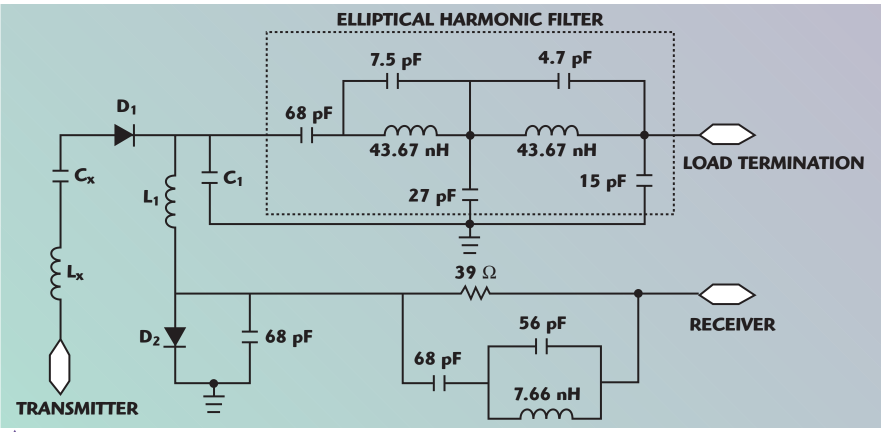


▲ *Fig. 1 Theoretical parallel-circuit class E topology combining a reactance compensation network.*

The parallel-circuit class-E PA provides the condition that high currents and high voltages do not occur simultaneously. This condition minimizes the power dissipation within the device. The load network consists of a parallel inductance L, a parallel capacitance C, a series L_0C_0 resonant circuit tuned at the fundamental and a load R,[3] as shown in ***Figure 1***.



▲ *Fig. 2 Practical broadband class E parallel circuit combined with a reactance compensation technique.*



▲ *Fig. 3 Schematic of the series LC network with an elliptical harmonic filter and RF switching of the 135 to 175 MHz range.*

The shunt LC-circuit provides a constant load phase angle relative to the device output terminals.[3] An additional low-pass matching section, transforming the impedance from load R to a 50 Ω termination, can be typically implemented to increase the harmonic suppression. ***Figure 2*** shows a broadband (135 to 175 MHz) class-E parallel-circuit load network, combining a reactance compensation technique with an additional three-section matching network to deliver a power of 8 W, using a 7.2 V supply voltage.

The harmonic suppression, in two-way radio applications, needs a typical 73 dBc reduction of the second harmonic at VHF and UHF frequencies and, typically, higher harmonics are not an issue. An additional high-Q roll-off harmonic filter is definitely required to provide high suppression of the second and other harmonic levels. A high-Q roll-off elliptical harmonic filter is chosen. However, in two-way radio applications, an RF switch is necessary between receiver and transmitter modes. ***Figure 3*** shows an optimized elliptical harmonic filter and RF switch for the broadband frequency range of 135 to 175 MHz.

A pin diode switch (D1 and D2) is turned on by an external DC source and provides low impedance. L_1 and C_1 are tuned for adequate receiver isolation. With the cascade of an elliptical harmonic filter and RF switch, the class-E switching action for broadband frequency (135 to 175 MHz) was not fulfilled. The combination of the elliptical harmonic filter and RF switch section was built and the S-parameters of the two-port network were measured with an HP6788 network analyzer. The measured results showed that the real input impedance was approximately 46 Ω and the imaginary input impedance moved from inductance to capacitance with a high variation of 20 Ω across the broadband range, as shown in ***Figure 4***. In the design, the transformation of the output impedance load network from R to 46 Ω load impedance is required.

A series L_xC_x compensation network is introduced to flatten the phase angle, as shown in ***Figure 5***. The measured results showed that, with the series L_xC_x, the imaginary variation is minimized to 5 Ω across the bandwidth without affecting the real

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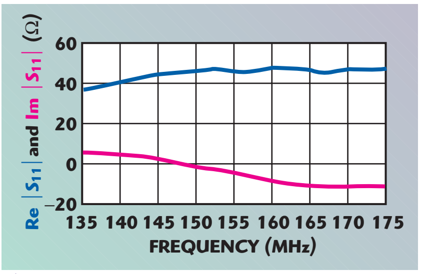
part. With the series L_xC_x, the imaginary impedance behaves as a capacitance across the bandwidth. This way, the three-section matching network can absorb the variation. The measured passband insertion loss and harmonic attenuation of the elliptical harmonic filter and RF switch with series L_xC_x are shown in ***Figure 6***. The insertion loss across the bandwidth is 1.2 dB and the second-harmonic attenuation is greater than 45 dB.

IMPLEMENTATION AND EXPERIMENTAL RESULTS VALIDATION

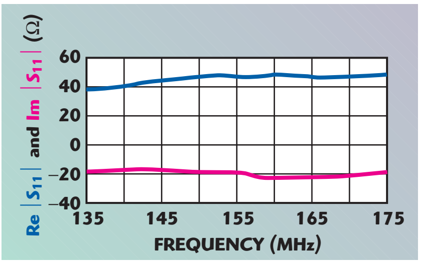
A prototype board of the class-E final PA, using a parallel-circuit combining reactance compensation together with a series L_xC_x network, an elliptical harmonic filter and RF switch was designed and fabricated. The first- and second-stage PAs were built as well. The PCB (FR-4 material) has a dielectric ε_p = 4.5 and thickness h = 14 mils. A heatsink is mounted at the bottom of the final PA in the PCB.

Since the insertion loss of the elliptical harmonic filter and RF switch is 1.2 dB across the bandwidth, the final PA must compensate the losses to deliver 6.5 W of power at the load termination. The final PA is delivering 8 W with a supply voltage of 7.2 V. A typical gate voltage of 1.8 to 1.9 V has been applied to the gate of the final PA. In order to provide sufficient RF input to the final PA with fast rise and fall times across the bandwidth, the gate voltage of the second-stage PA can be adjusted (a gate voltage of 2.1 to 2.4 V is applied).

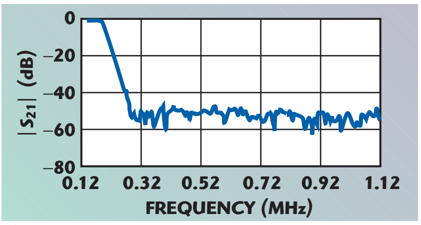
The measured second- and third-harmonic levels across the bandwidth are shown in ***Figure 7***. The measured output power at the load termination and the final PA efficiency are shown in ***Figure 8***. A final PA efficiency of 70 percent at an operating power of 6.5 W was demonstrated



▲ *Fig. 4 Measured real and imaginary input impedance of the elliptical harmonic filter without the series Lx Cx network.*

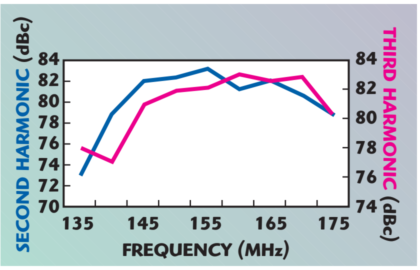


▲ *Fig. 5 Measured input impedance of the elliptical harmonic filter with the series Lx Cx network.*

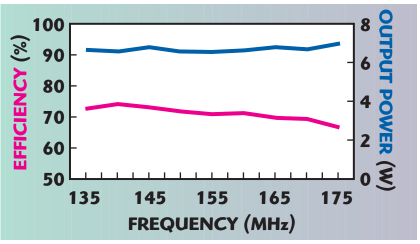


▲ *Fig. 6 Measured passband insertion loss and harmonic attenuation of the elliptical harmonic filter with RF switch and series compensation.*

across the bandwidth (135 to 175 MHz) and a second-harmonic attenuation of –73 dBc was achieved across the bandwidth. A third-harmonic attenuation greater than –75 dBc was also achieved across the bandwidth.



▲ *Fig. 7 Measured second- and third-harmonic levels across the frequency band.*



▲ *Fig. 8 Measured output power and efficiency across the frequency band.*

CONCLUSION

High harmonic suppression while maintaining a class-E condition for a broad frequency range is achieved with a new type of load network circuitry. A broadband parallel-circuit class-E network, cascaded with a series L_xC_x and a high-Q roll-off elliptical harmonic filter was used for this implementation. The novel L_x C_x circuit compensates the large reactance variation of the elliptical harmonic filter, which enables class-E PA operation. An efficiency of 70 percent and a second-harmonic suppression of –73 dBc was measured across a wide bandwidth (135 to 175 MHz) with an operating power of 6.5 W. The efficiency and harmonic results are the highest reported so far at VHF frequencies with a 40 MHz bandwidth and a supply voltage of 7.2 V. ■

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T82-50	2-8	50	47
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CMC500	80-1000	500	57
CMC1000	80-1000	1000	60
SMX Series • .01-1000 MHz			
SMX100	.01-1000	100	50
SMX200	.01-1000	200	53
SMX500	.01-1000	500	57
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SMV500	500-1000	500	57

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The speed to market is critical to the success of any RFIC company. With more and more wireless standards proposed and adopted, the design cycles are actually getting shorter. It puts a lot of pressure on the RFIC team to come up with the design faster. In the digital CMOS design, the basic building blocks can be scaled based on the process. In RFIC CMOS design, each transceiver has to be tailored to the specific application because RFIC is heavily frequency dependent. Even though the new RFIC cannot simply be scaled down as the digital IC, the basic topologies can be reused extensively. After examining hundreds of transceiver RFICs and their building blocks, it is easy to see what the dominant topology is for each building block in a typical RFIC. In this article, the most widely used topologies of each building block are presented. Those building blocks can be the starting point for any generic RFIC design. They need to be optimized for the specific application. For example, the matching circuits need to be designed according to the frequency, even though the topology is the same.

For an RFIC company, a modular RFIC standard cell approach will shorten the design cycle and reduce the design risks.

A generic wireless transceiver consists of three main blocks: a transmitter (TX), a receiver (RX) and a synthesizer. The major building blocks in a direct conversion TX are a baseband filter, a modulator, the RF filter, an automatic gain control (AGC) circuit, a power amplifier (PA) and the power detector loop. The modulator (MOD), the AGC and the PA are discussed extensively in this article. A direct conversion RX normally consists of a RF SAW, a low noise amplifier (LNA), an image rejection SAW, a demodulator (DEMOD), an AGC and a RX baseband filter. The DEMOD and the AGC topology in a RX are similar to the ones used in a TX with minimal modifications. They will be covered in the TX MOD and AGC discussion. The synthesizer is typically made up of a voltage-controlled oscillator (VCO), a phase/frequency detector (PFD), a charge pump (CP), a frequency divider and a digital logic control block. For the synthesizer discussion, all but the digital logic control will be discussed.

This article starts with the discussion on the LNA. The PA is next. The MOD/DEMOD and the AGC discussion are shared for a TX/RX. The article ends with the discussion on the synthesizer.

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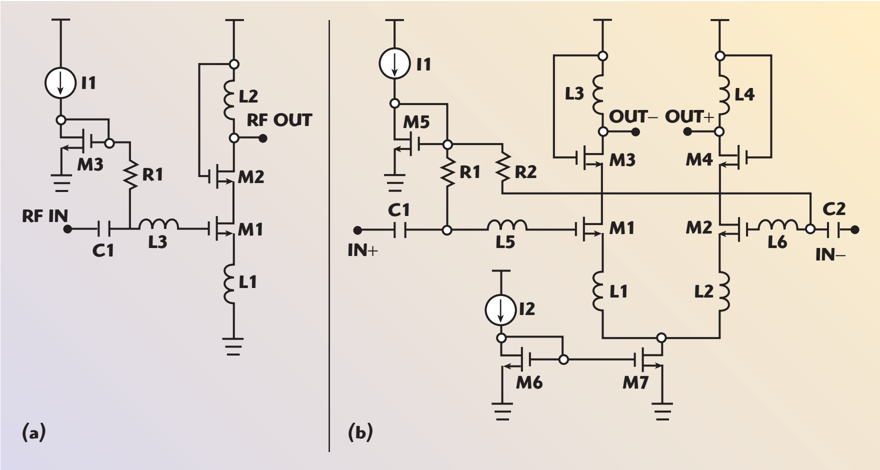
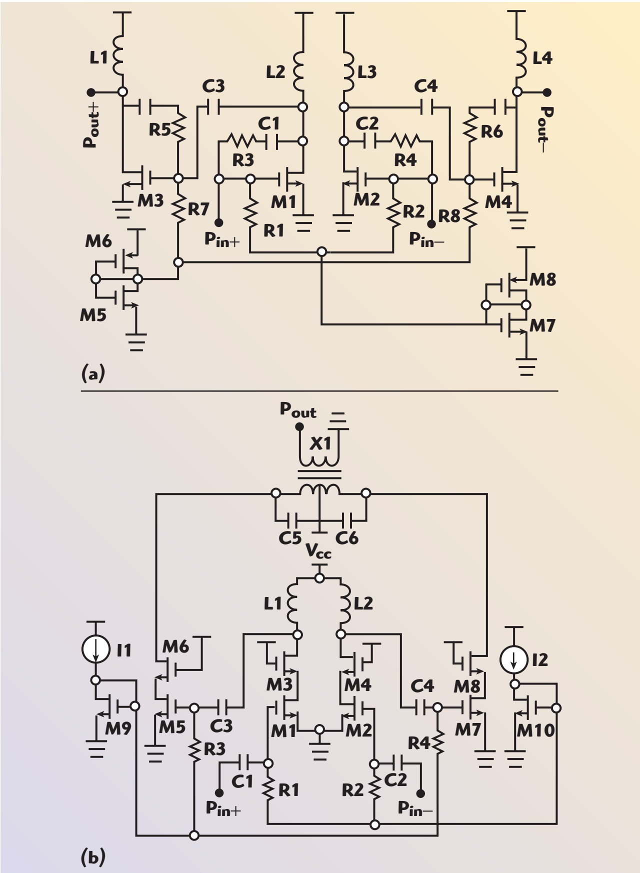


▲ *Fig. 1 CMOS LNAs; (a) single-ended and (b) differential versions.*

The LNA's performance directly impacts the receiver performance, especially the RX sensitivity. Gain, noise figure (NF), input third-order intercept (IIP3), stability and I/O matching are important design parameters. It is difficult to achieve the best performance in one parameter without sacrificing another. The classical tradeoff is gain versus NF. A single-ended and a differential version of the popular LNA designs are shown in ***Figure 1***.

The Miller effect increases the parasitic capacitance at the gate of the LNA. A cascode topology reduces this effect by stacking a common gate (CG) stage on a common source (CS) gain stage. The input and output port are better isolated to reduce the parasitic capacitance between the gate and drain. In the single-ended version, M1 is the CS stage while M2 is the CG stage. L1 is the degenerated feedback element to bring the input NF circle and the gain circle closer. Thus, a compromise between NF and gain can be reached. L3 is the input matching element. C1 is the input DC block. The LNA is biased in a current mirror configuration with M3, I1 and R1. In a highly integrated IC circuit, the common mode noise is a major problem. A differential circuit is often used to combat this problem. The differential LNA can be considered as two single-ended LNAs with the tail bias current. The tail current out of M7 is important to reduce the common mode noise. Without it, there will not be enough common mode degenerated resistance to reduce the common mode noise. The bias current is set by the current mirror consisting of M6 and M7. It, in turn, sets the Vgs of M1 and M2. The drain voltage of M7 is set by the current mirror biasing at the gate of the M1 and M2.



▲ *Fig. 2 CMOS PA; (a) differential and (b) cascode differential.*

The PA is the next block on deck. The CMOS PA has gradually replaced the HBT or GaAs FET in the low to medium power applications like Bluetooth, WLAN, etc. Its performance is inferior, compared to the HBT or the GaAs FET, but the CMOS PA can be easily and cheaply integrated in the CMOS transceiver. Many design imperfections can be corrected by the baseband digital processor.

The PA design is similar to a LNA or a general-purpose gain block, but its emphasis is on output power, linearity and efficiency. The design starts at the output port where the output power contours of the device are characterized. Once the output termination is determined, the matching circuit is designed the same way as the other RF building block's matching circuits. A two-stage common source FET with RC feedback is shown in ***Figure 2***. M1 and M2 are the driver stages. M3 and M4 are the output power stage. The RC feedback from the drain to the gate stabilizes the transistors at the high frequency. The feedback also broadens the bandwidth of the PA. M5 and M6 and M7 and M8 are the active resistive dividers to bias the driver stage and the power stage, respectively. The cascode PA design is popular as well. Topology wise, it is similar to the cascode LNA. M1 through M4 are the driver stage, while M5 to M8 complete the power stage. An integrated balun is utilized in this design since the most widely used antennas in wireless devices are single ended.

The DEMOD is used to convert an RF signal to baseband in a direct conversion receiver or to a low IF signal in a Very Low IF (VLIF) RX. Since most modern wireless devices require both I and Q channels, two double-balanced mixers (DBM) are needed in the DEMOD. The implementation is illustrated in ***Figure 3***. M1 to M6 is the DBM for the I channel. M11 to M16 is the DBM for the Q channel. Since the same DBM is used for both I and Q channels, only one DBM is discussed in detail. The

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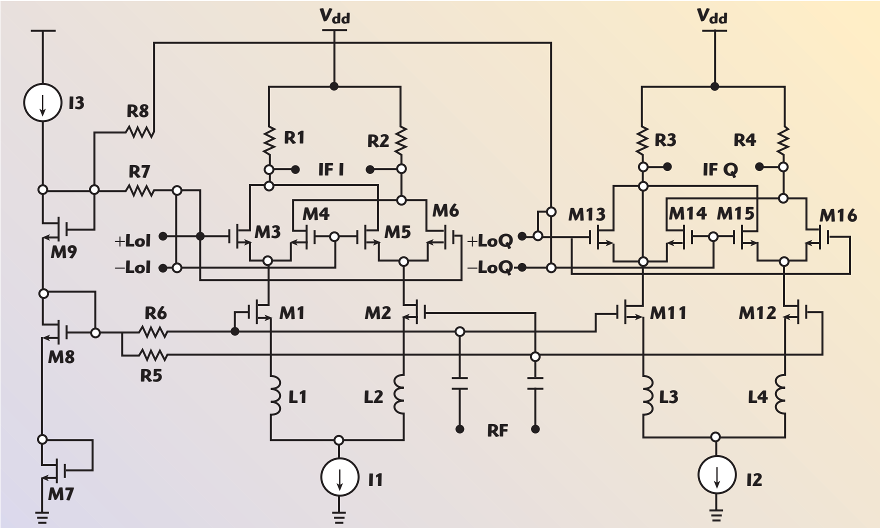
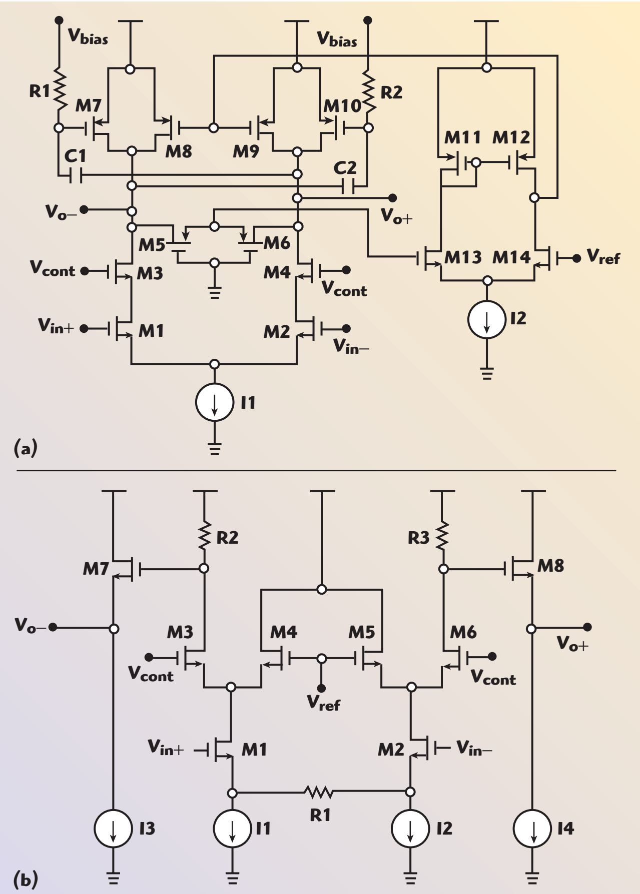


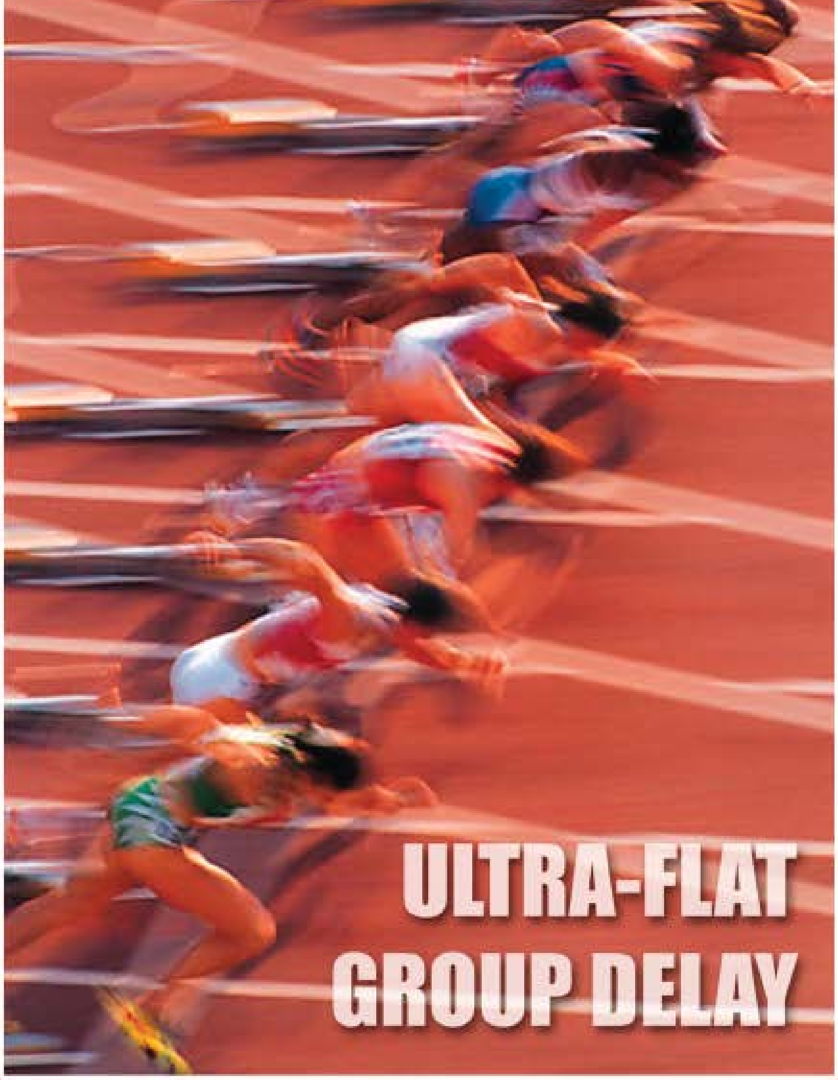
▲ *Fig. 3 CMOS DEMOD.*

incoming RF signal is amplified first by the gain stage such as M1 and M2. M3 to M6 are the switching FETs. It fundamentally serves the purpose of a multiply operation. In half the cycle, M3 and M6 are on. The local oscillator (LO) and RF are essentially multiplied in phase. In the other half of the cycle, M4 and M5 are turned on to reverse the polarity of the output signal. The output loads are implemented with resistors R1 to R4. The degenerated feedback inductors L1 to L4 help the IIP3 performance. The biasing is done with the diode connected FETs. The MOD's topology is similar to the DEMOD. The RF and IF positions need to be swapped. The signal-combining network needs to be swapped as well.

The same AGC circuit can be used in either a TX or a RX. There are many AGC topologies to choose from. The variable transconductance and the variable biasing AGC are the most popular for the high frequency operation. They are shown in ***Figure 4***. The variable transconductance circuit is based on the principle that the transconductance of the FET changes as the FET goes from a saturation mode to a triode mode. Thus, the gain is varied. M1 to M4 can be considered as the typical differential cascode gain stage. The difference is that V_{cont} is applied at the gates of M3 and M4. As V_{cont} decreases, the drain voltage at M1 and M2 drops. Eventually M1 and M2 enter the triode region. M7 to M10 are the current source type load. Thus, a common mode feedback (CMFB) is needed to make sure the current source matches with the current sink I1. The common mode voltage is sampled at the drain of M3 and M4 with M5 and M6. M5 and M6 can be considered as two large value resistors. They have the same value. The sampled voltage is fed to a comparator (M11 to M14). The reference is fed to one in-



▲ *Fig. 4 AGC; (a) variable transconductance and (b) variable biasing.*



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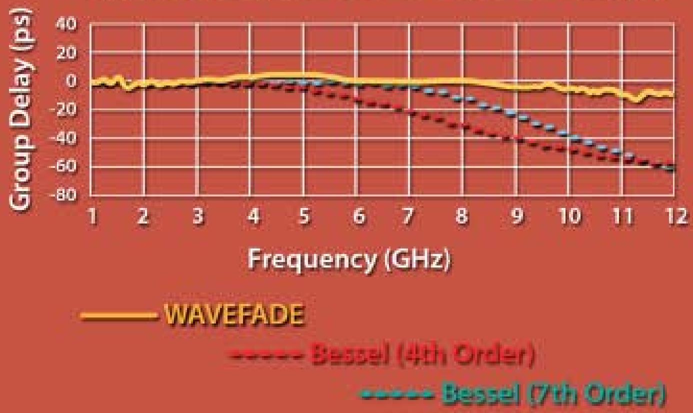

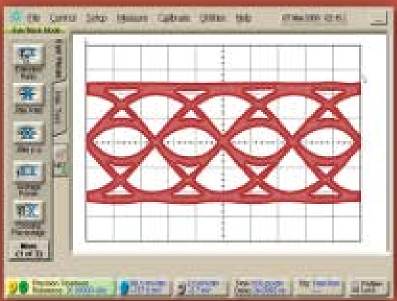

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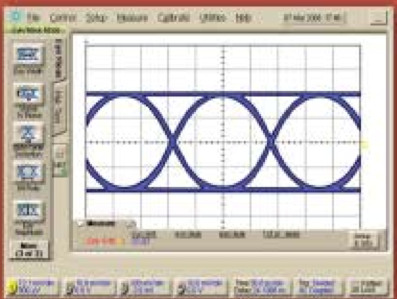

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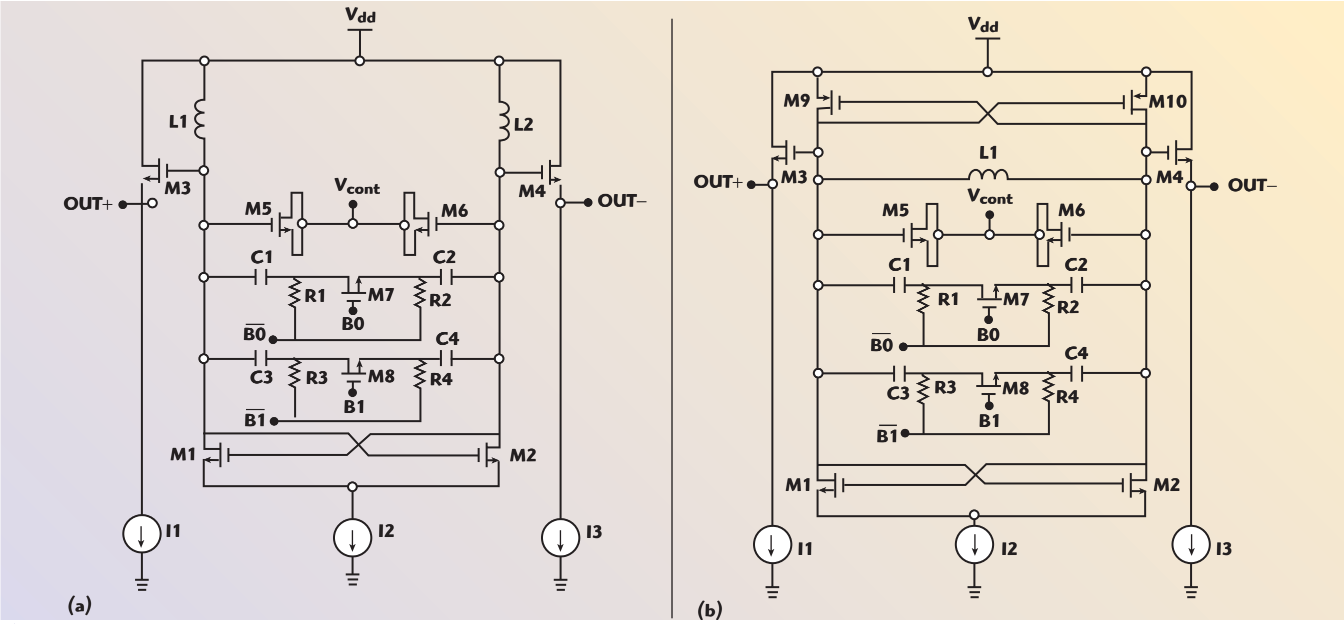


Fig. 5 VCOs; (a) NMOS only and (b) complementary CMOS.

put M14, while the sampled common mode is fed to the other input M13. The error voltage is used to control the bias current out of the current source M8 and M9. M8 and M9 are the current bleeding path. The closed feedback loop ensures correct bias current follows with the desired reference voltage.

The variable current AGC is based on the idea that the tranconductance of a FET changes with the bias current. By varying the bias current in the FET, the AGC can be accomplished. M1 and M2 are the input gain buffer. M4 and M5 are the current bleeding path. If V_{cont} is larger than the V_{ref}, more bias current flows through M3 and M6. In this mode, a high gain is expected. When V_{cont} is reduced, more bias current flows through M4 and M5, and the gain is thus reduced. M7 and M8 are the output emitter follower buffers to reduce the output impedance. R1 is the degenerated resistor to improve the linearity of the AGC.

The CMOS VCO is based on the negative resistance theory. Two approaches are shown in ***Figure 5***. In the NMOS only version, by cross coupling M1 and M2, a positive feedback is created in the circuit. Looking into the gate of M1 and M2, a negative resistance can be expected. The operating frequency is set by the tank circuit resonance. L1

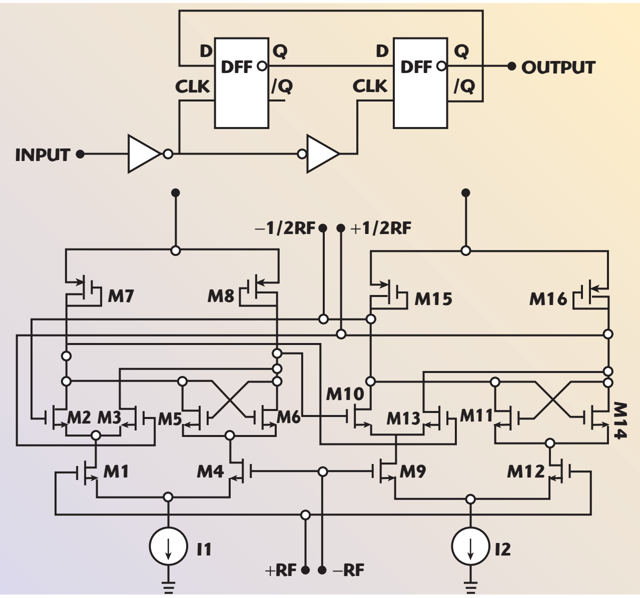


Fig. 6 Frequency divider.

and L2 provide the inductance part. The frequency can be tuned coarsely by the capacitance banks and fine-tuned with the varactor capacitors. In the circuit shown, two-bit, four-state capacitance banks are used. More banks can be added if a larger process variation is expected. By turning M7 and M8 on and off, more or less total capacitance can be expected in the resonator tank. The varactor is implemented with the FETs M5 and M6 by tying the source and drain. M3 and M4 are the output buffers.

In the complementary CMOS VCO, the key change is to add a cross-coupled PMOS pair. By adding a PMOS pair, two more elements are added to contribute to the negative resistance while the bias current remains the same. It is more power efficient.

The high-speed frequency divider or a prescaler are typically implemented with a Johnson counter or a divide-by-two building block. Its block diagram and detailed implementation are shown in ***Figure 6***. The Johnson counter consists of a master DFF (D-type flip-flop), a slave DFF and a couple of inverters. In 50 percent of the input duty cycle, the master DFF is on. For the rest of the 50 percent duty cycle, the slave DFF is on. The divided-by-two outputs are taken from the outputs of both DFF. The outputs are 90° phase shifted by the nature of the Johnson counter. Since the master and slave DFFs are identical, only the master DFF is discussed in detail. A DFF is based on the emitter-coupled logic. This logic works by steering the bias current. Thus, it is a faster logic compared to the CMOS logic. M1 and M4 are the input buffer. M5 and M6 are the latch element. M2 and M3 can be considered as the mode switch to decide to load a new input value or

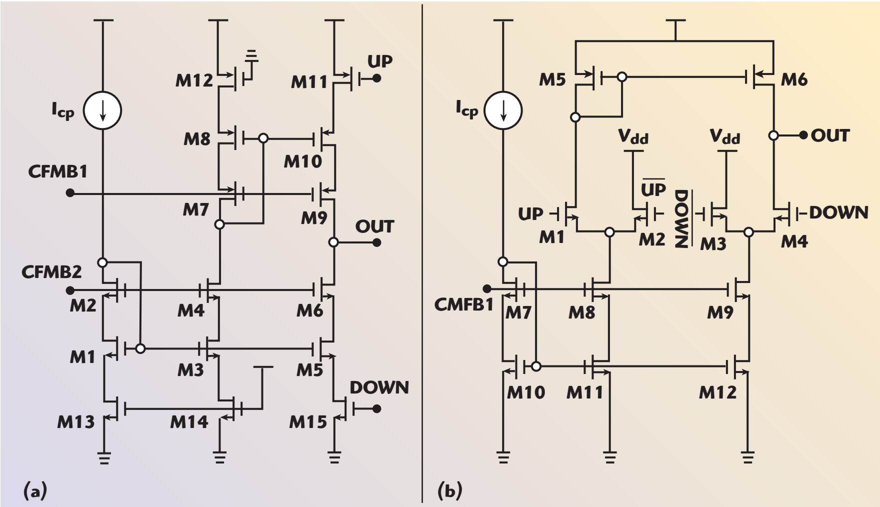
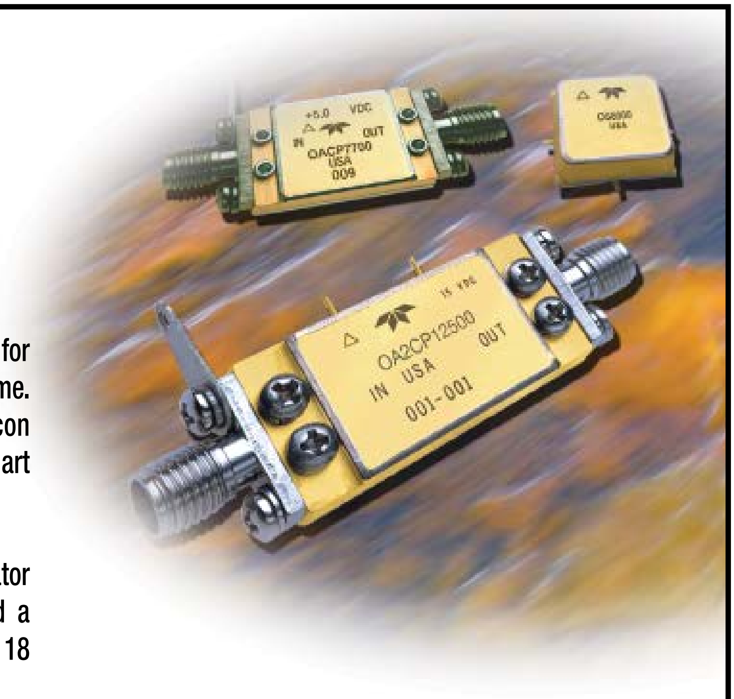


Fig. 7 CMOS charge pump; (a) switching at the source and (b) current steering CP.

Top Performers:
85 MHz to 18 GHz
VCOs

Teledyne Cougar's VCO circuits are designed for best overall phase noise and fastest settling time. Using silicon bipolar transistors and silicon varactor diodes make Cougar VCOs the smart choice for performance challenges.

Cougar's designs use a fundamental oscillator approach for the 85 MHz to 9 GHz VCOs and a doubling oscillator approach for the 9 GHz to 18 GHz VCOs.



Model	Freq. Range (MHz)	Tuning Voltage Range (Volts)	Output Power/ Variation (dBm/ ±dB)	Typical Phase Noise Offset at 10kHz/100kHz (dBc/Hz)	Nominal Modulation Sensitivity Min.-Max. (MHz/V)	Typical Harmonic Suppression (dBc)	D.C. Bias Voltage (Volts)	D.C. Bias Current (mA)
Oscillator with internal MMIC amplifier available in SMTO-8 or CougarPak™								
OAS5100	4300-5100	0-15	13.0/2.0	-84/-108	50-85	-22	5.0	94
OAS7700	5700-7700	0-15	10.0/2.0	-75/-100	70-250	-30	5.0	95
OAS8900	6900-8900	0-15	10.0/2.0	-70/-95	100-270	-30	5.0	95
Oscillator available in SMTO-8 or CougarPak™								
OS6700	5400-6700	0-15	0/2.0	-75/-100	80-180	-17	5.0	25
OS7700	5700-7700	0-15	2.0/2.0	-75/-100	70-250	-17	5.0	25
OS8900	6900-8900	0-15	1.0/2.0	-70/-95	100-270	-25	5.0	24
Oscillator available in TO-8, SMTO-8 or CougarPak™								
OC150	85-150	0-20	10.0/2.0	-100/-120	2-8	-8	15.0	35
OC1850	1250-1850	0-15	8.0/2.0	-83/-111	20-70	-12	15.0	37
OC4500	3500-4500	0-15	8.0/2.0	-75/-100	50-150	-10	15.0	60
Oscillator, Amp, Filter and Voltage Regulator in 2- and 3-Stage CougarPak™								
OA2CP1001	500-1000	0-(-12)	15.0/2.0	-75/-105	30-60	-15	15.0	250
OA2CP12500	9000-12500	0-(-12)	15.0/2.0	-65/-95	150-450	-25	15.0	250
OA3CP18001	12000-18000	0-(-12)	15.0/2.0	-55/-85	150-750	-15	15.0	350

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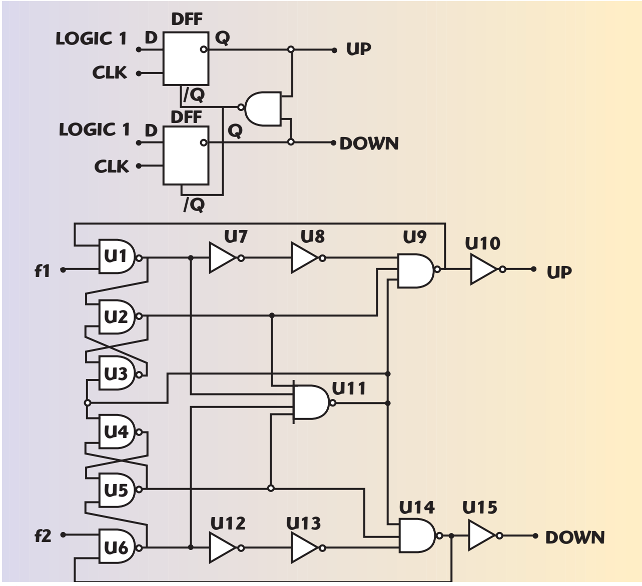

Fig. 8 CMOS PFD.

to keep the current value. M7 and M8 are the diode tied load. A resistor load can be used as well.

The charge pump is used in the synthesizer to source or sink current from the external loop filter. The basic idea is to add a switch in both the current source and current sink paths. The switch can be added at the gate, source or drain, as shown in ***Figure 7***. To reduce the effect of the charge sharing, clock feed-through and charge injection, switching at the source is the best choice because the switch is relatively isolated from the output. I_{cp} is the reference current for both the current sink and the current source. For the current sink path, I_{cp} is mirrored into the M6 and M5 path through the modified stacked current mirror (M1, M2, M5 and M6). The advantage of this current mirror is the low supply voltage requirement. M15 is the source switch for the current sink. M13 is a dummy FET to ensure that M1 and M5 have the same DC source voltage. For the current source path, I_{cp} is first mirrored by a modified stacked current mirror (M1, M2, M3 and M4). Then it is mirrored again by the PMOS current mirror (M7, M8, M9 and M10). M11 is the switch for the current source.

To reduce the mismatch problem and to increase the speed of the CP, a current steering CP is often used. The switches are implemented with a current steering pair (M1 and M2, M3 and M4). I_{cp} is mirrored to the current steering pair via the modified stacked current mirror. To source the current, UP is logic H and /UP is logic low. I_{cp} goes through the path of M1 and then it reaches the output via another current mirror (M5 and M6). To sink the current, the DOWN control pair is activated. Because the current is steered instead of charging/discharging, it will be faster than the switching at the source approach. However, it will burn more power since the current path is not really shut off when it is not used.

The UP/DOWN control signals used in CP are generated by the PFD. The two DFFs implementation has been a workhorse for a long time. Its block diagram and the circuit implementation are shown in ***Figure 8***. The upper and lower DFFs generate UP and DOWN signals, respectively. When f1 arrives first at the upper DFF, the UP sig-

nal is generated to turn on the current source. When f2 arrives at the the lower DFF, its output is logic H. When both UP and DOWN are logic H, NAND gate resets both DFFs to output logic L. Thus, the phase difference between the two signals are detected and used to turn on either the current sink or current source of the CP.

Each major building block in a modern wireless transceiver is discussed in this article. The circuits are generic enough to be adopted in most wireless applications. It will be offered as a good starting point for the next RFIC design. For an RFIC company, a modular RFIC standard cell approach will shorten the design cycle and reduce the design risks. For a foundry company, a portfolio of RFIC IPs will add more value to its customers. The RFIC designs no longer need to start with a blank paper. The modular standard cell approach will make future RFIC designs easier to tackle. ■

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Louis Fan Fei *received his BEE and MSEE degrees from Georgia Tech in 1996 and 1998, respectively. He worked on microwave instrument circuits for HP/Agilent in Colorado Springs, CO, in the summer of 1997 and on WLAN and wireless local loop circuits at Lucent/Agere Systems from 1998 to 2003. He is now an RF engineer at Garmin International, where he designs GPS receivers.*

A FUNDAMENTAL WIDEBAND LOW NOISE VCO SERIES

EXECUTIVE INTERVIEW SERIES

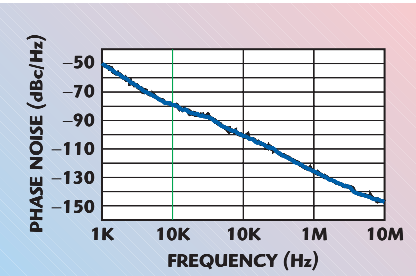
MWJ SPEAKS WITH OLLE WESTBLOM, MANAGING DIRECTOR, SIVERS IMA AB.

VISIT WWW.MWJOURNAL.COM TO READ THIS IN-DEPTH INTERVIEW.

Signal sources play a key role in many microwave systems, with the quality of the microwave signal being particularly critical as the complexity of modulation increases. During its operation over nearly five decades, Sivers IMA has been responsible for the production of voltage-controlled oscillators (VCO), dielectric resonator oscillators (DRO), yttrium iron garnets (YIG) and subsystems. Over the last decade, however, the emphasis has been on VCOs as the basis for signal source and frequency modulated continuous wave (FMCW) products, with fundamental wideband low noise VCOs steadily increasing in popularity.

The company's latest introduction is the VO3280 series, which is a drop-in fundamental wideband low noise VCO intended for any application where size and performance is important. It has evolved from the company's VO3260 series, with the new integrated microwave monolithic integrated circuit (MMIC) design enabling high repeatability and thus reducing the work of the system integrator. Although wideband and low noise are optimized, the series also achieves low current consumption and high output power.

Fig. 1 Typical phase noise characteristics for the VO3280 VCO family.



A narrow band VCO can take advantage of high Q in the resonator circuit and thereby achieve a low phase noise. Normally, when designing a wideband VCO, compromises often need to be made between the tuning range and the phase noise capabilities. With the VO3280, however, Sivers IMA has designed a VCO series combining both low phase noise and a broad tuning range. ***Figure 1*** shows the phase noise characteristics of the series.

The VCO series is designed for wire or ribbon bonding, but is also available in various packages with SMA connectors and solder pins. It weighs 2.5 g and has ruggedized design incorporating a mixture of ceramics and metal—the carrier and lid are made from Kovar, gold plated over a barrier of nickel, while the substrate is ceramic and the connections are gold plated on tantalum nitride. This robust-

SIVERS IMA AB
Kista, Sweden

MILLIMETER WAVE RECTANGULAR TE_{10} WAVEGUIDE INFORMATION

WG Band	Waveguide Frequency Range (GHz)	Wavelength Range λo (mil)	Wavelength Range λo (mm)	Guide Wavelength Range ($\lambda g/\lambda o$)	Waveguide Impedance Range (Ω)	TE_{10} Cutoff Freq (GHz)	TE_{10} Cutoff λc (mil)	TE_{10} Cutoff λc (mm)	Internal Dimensions (mils)	Internal Dimensions (mm)
WR-28	26.5 - 40.0	445.4 - 295.1	11.313 - 7.495	1.650 - 1.177	621.9 - 443.6	21.1	560.0	14.22	280.0 x 140.0	7.112 x 3.556
WR-22	33.0 - 50.0	357.7 - 236.1	9.085 - 5.996	1.661 - 1.177	626.0 - 443.6	26.3	448.0	11.38	224.0 x 112.0	5.690 x 2.845
WR-19	40.0 - 60.0	295.1 - 196.7	7.495 - 4.997	1.613 - 1.173	608.3 - 442.4	31.4	376.0	9.55	188.0 x 94.0	4.775 x 2.388
WR-15	50.0 - 75.0	236.1 - 157.4	5.996 - 3.997	1.657 - 1.181	624.8 - 445.1	39.9	296.0	7.52	148.0 x 74.0	3.759 x 1.880
WR-12	60.0 - 90.0	196.7 - 131.1	4.997 - 3.331	1.690 - 1.186	637.2 - 447.1	48.4	244.0	6.20	122.0 x 61.0	3.099 x 1.549
WR-10	75.0 - 110.0	157.4 - 107.3	3.997 - 2.725	1.620 - 1.185	610.9 - 446.7	59.0	200.0	5.08	100.0 x 50.0	2.50 x 1.270
WR-08	90.0 - 140.0	131.1 - 84.3	3.331 - 2.141	1.746 - 1.177	658.1 - 443.6	73.8	160.0	4.06	80.0 x 40.0	2.032 x 1.016
WR-06	110.0 - 170.0	107.3 - 69.4	2.725 - 1.763	1.771 - 1.183	667.7 - 445.9	90.8	130.0	3.30	65.0 x 32.5	1.651 x 0.826
WR-05	140.0 - 220.0	84.3 - 53.6	2.141 - 1.363	1.777 - 1.176	669.7 - 443.3	115.7	102.0	2.59	51.0 x 25.5	1.295 x 0.648
WR-04	170.0 - 260.0	69.4 - 45.4	1.763 - 1.153	1.695 - 1.177	638.8 - 443.9	137.2	86.0	2.18	43.0 x 21.5	1.092 x 0.546
WR-03	220.0 - 325.0	53.6 - 36.3	1.363 - 0.922	1.627 - 1.183	613.5 - 445.9	173.6	68.0	1.73	34.0 x 17.0	0.864 x 0.432
WR-02.8	260.0 - 400.0	45.4 - 29.5	1.153 - 0.749	1.708 - 1.177	643.8 - 443.6	210.8	56.0	1.42	28.0 x 14.0	0.711 x 0.356
WR-02.2	325.0 - 500.0	36.3 - 23.6	0.922 - 0.600	1.771 - 1.185	667.7 - 446.7	268.2	44.0	1.12	22.0 x 11.0	0.559 x 0.279
WR-01.9	400.0 - 600.0	29.5 - 19.7	0.749 - 0.500	1.587 - 1.169	598.3 - 440.6	310.6	38.0	0.97	19.0 x 9.5	0.483 x 0.241
WR-01.5	500.0 - 750.0	23.6 - 15.7	0.600 - 0.400	1.620 - 1.175	610.9 - 442.8	393.4	30.0	0.76	15.0 x 7.5	0.381 x 0.191
WR-01.2	600.0 - 900.0	19.7 - 13.1	0.500 - 0.333	1.746 - 1.194	658.1 - 450.1	491.8	24.0	0.61	12.0 x 6.0	0.305 x 0.152
WR-01.0	750.0 - 1100.0	15.7 - 10.7	0.400 - 0.273	1.620 - 1.185	610.9 - 446.7	590.1	20.0	0.51	10.0 x 5.0	0.254 x 0.127

TABLE I

THE VO3280 SERIES VCOs

Standard Frequency Band (GHz)	Drop-in Stripline	Drop-in Coplanar	SMA-SMA Housing	SMA-Solder Pin Housing
2 to 3	VO3280S/01	VO3284S/01	VO4280S/01	VO4283S/01
3 to 5	VO3280S/00	VO3284S/00	VO4280S/00	VO4283S/00
5 to 8.4	VO3280C/00	VO3284C/00	VO4280C/00	VO4283C/00
8.4 to 13.5	VO3280X/00	VO3284X/00	VO4280X/00	VO4283X/00
13.5 to 20	VO3280P/00	VO3284P/00	VO4280P/00	VO4283P/00
20 to 25.5	VO3280K/00	VO3284K/00	VO4280K/00	VO4283K/00

ness makes it suitable for tough environments such as in airborne, military and space applications.

With only six units, which are listed in ***Table 1***, the VO3280 series covers the frequency range from 2 to 25.5 GHz, with an output power of 15 dBm, while the VO3262 series provides +25 dBm output power and is recommended for use in high level mixer applications. The VO3280 and VO3284 are just 2.5 g (0.035 oz) units. The VO4280 is a radiated and mechanical protected version in a SMA house and the VO4283 series is a basic variant of a SMA connected version.

BIAS

Figure 2 is a block diagram of the VO3280. Even though it is fitted with an internal regulator, Sivers IMA recommends a low noise power supply and proper bias filtering to obtain a low noise RF signal. It is important for users to avoid using a switched (switch-mode) power supply directly on the bias input, or be extra careful with the bias filtering and a linear power supply. A 100 nF capacitor close to the bias connection is preferred.

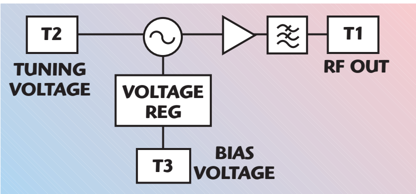


Fig. 2 Block diagram of the VO3280 series.

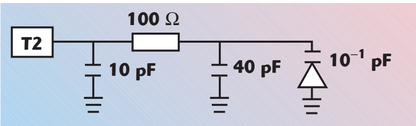


Fig. 3 Tuning input interface of the VO3280 series.

TUNING

The oscillator's tuning input voltage is between 0 and +20 V and normally the tuning current is very low. In order to keep the slew rate (or tuning speed) as high as possible, the filtering is kept to a minimum. There can be some RF leakage on the tune and bias terminals. On demand the modulation bandwidth can be adjusted to > 100 MHz.

S-band and C-band oscillators are temperature-compensated bipolar oscillators that will not oscillate below 2 V tuning voltage. If the oscillator is required to run continuously, such as the idle in a phase-locked loop (PLL), the tuning voltage should be kept above 2 V in order to keep the PLL locked. ***Figure 3*** shows the tuning interface.

RF OUTPUT

The RF output is a 250 µm wide microstripline, where Z_o = 50 Ω and is AC coupled with a 10 pF capacitor. A built-in low pass filter reduces the harmonic to better than –25 dBc and the VO3280 series provides up to +18 dBm output power, as shown in ***Figure 4***.

Importantly too, a coplanar ground-signal-ground (GSG) output is available and since the VCOs are fundamental, any sub harmonics are eliminated. Harmonics are suppressed by a built-in low pass filter. The VCOs show very good phase noise performance, considering the tuning range for some models is as high as 50 percent. Bipolar oscillator transistors are used up to 10 GHz in order to obtain the best phase noise performance possible and above 10 GHz GaAs FET transistors are used.

Due to the advanced design, even the FET VCO is powered by a single +15 V power supply and the built-in linear voltage regulator ensures low

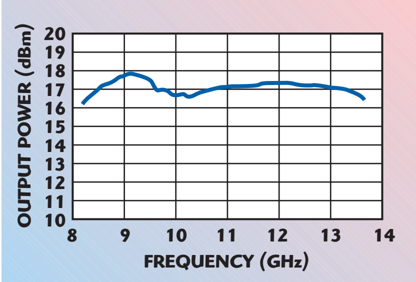


Fig. 4 The VO3280 series' output power characteristics.

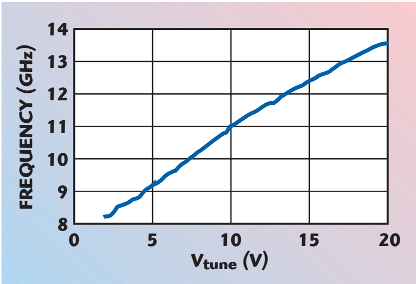


Fig. 5 Tuning characteristics with a tuning voltage of 0 to 20 V.

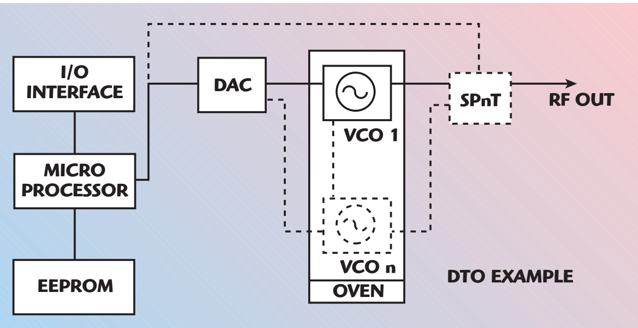


Fig. 6 Block diagram showing an example of a DTO.

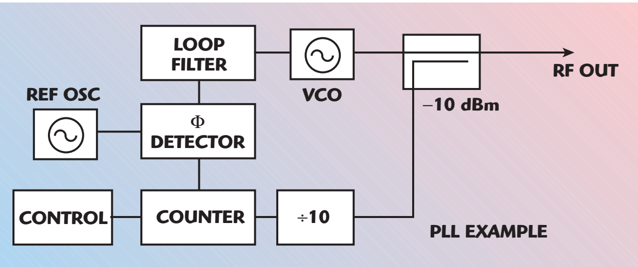


Fig. 7 Block diagram showing an example of a PLL.

pushing, making application design easier. The tuning voltage is between 0 and +20 V, and the hyperabrupt varactor oscillator design gives an excellent tuning ratio, as shown in ***Figure 5***. Every oscillator is individually tuned and the performance is optimized in-house.

APPLICATIONS

One of the main benefits of the VO3280 series is its wideband capabilities. Instead of designing an oscillator bank with control components and control logic, a wideband VCO can be used instead, simplifying the system design dramatically. These VCOs are used in countless applications. However, two are more common—digital tuned oscillators (DTO) and phased-locked loops.

DTO EXAMPLE

DTOs are often used in applications where the speed of frequency shift is critical, such as a local oscillator (LO) in set on receivers (SOR) or frequency hopping radar receivers. The table of calibrated frequency values is stored in an EEPROM. Temperature drift is eliminated due to the fact that the VCO is temperature stabilized either with an oven set for 55°C, or a Peltier thermoelectric module set for 35°C.

As an alternative, the VCO can be calibrated at several temperatures and an interpolated calibration table is stored in the EEPROM that is used together with a microprocessor. A temperature sensor in the VCO assembly gives the microprocessor the current temperature. In order to cover multi-octave frequency bands, several VCOs can be stacked and selected by the microprocessor (see ***Figure 6***).

PLL EXAMPLE

PLLs are used in applications where the stability and accuracy is essential for the function of the system, that is, as an LO in a coherent radar jammer. In this example (see ***Figure 7***), a sample of the output signal is divided digitally or down converted and fed into the phase/frequency detector. The detector compares the divided signal with the reference signal and a correction is given through the loop filter to the tuning circuit, adjusting the VCO until the divided and reference signal are in phase.

CONCLUSION

The VO3280 series features high bandwidth, low phase noise, high linearity, medium and high power, and an extended temperature range. It is a hermetically sealed, drop-in VCO available in standard and custom housings. Customized frequency ranges and power levels are offered and coplanar and stripline RF ports are available. Every unit is individually tested and its ruggedness means that it is suitable for tough environments. Additional information may be obtained via e-mail at sales@siversima.com.

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Kista, Sweden, +46-8-7036800;
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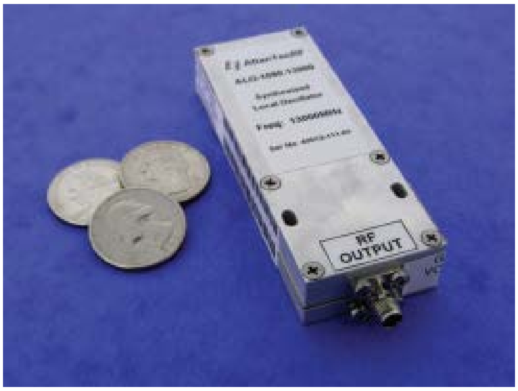


SYNTHESIZED LOCAL OSCILLATORS OFFER WIDE CHOICE OF OUTPUT FREQUENCY

In many commercial applications, particularly in the military and aerospace sphere, space is at a premium and component weight is a key consideration. In such situations, light weight and the flexibility afforded by modularity are key considerations, as is having a range of products to meet all requirements.

To meet these criteria, AtlanTecRF has developed the new ALO series of synthesized local oscillator modules that combine small size (33 × 88 × 16 mm) and light weight with a large range of output frequency. The modules are offered in frequency ranges with total coverage from 5 to 14.55 GHz.

These new devices complement the company's phase-locked oscillators in the APL series that utilize coaxial resonators up to 3 GHz and dielectric resonators (DRO) up to 14 GHz. These require a crystal-controlled reference oscillator in the 25 to 300 MHz range to be provided, either internally or externally, and as a result achieve very good phase noise. DRO-based oscillators are relatively narrow band in frequency due to the high Q of the DRO puck that has to be mechanically factory tuned and therefore has to be selected at the time of ordering, potentially lengthening the turnaround time.

PROGRAMMED TO SERVE

The required output frequency of the new ALO series is factory programmed to customer order with steps available in 500 kHz increments within the frequency range of each module. This is achieved via a group of pins in the power and alarm connector and can therefore be set to the required frequency very conveniently and quickly, thus shortening turnaround time. All models in the series share a common reference oscillator circuit based around a 20 MHz temperature-controlled crystal oscillator (TCXO).

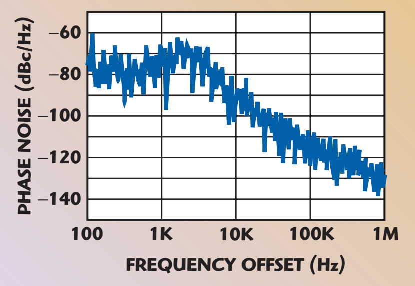
The new reference oscillator design provides a benefit in terms of temperature stability that is improved by a factor of more than two times and allows a greater operating temperature range over the older designs. Also, the frequency stability of the oscillators is very good as a result of the internal TCXO with a setting error of less than ±0.5 ppm and tem-

ATLANTECRF
Braintree, UK

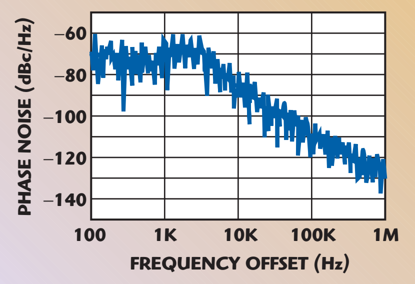
perature stability of less than ±2.5 ppm for the entire specified operating temperature range of –30° to +70°C.

PHASE NOISE

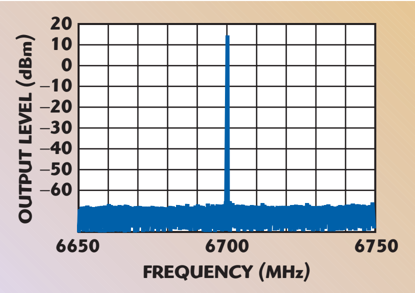
The synthesized oscillators have low phase noise at 100 kHz offset from the carrier of typically less than –105 dBc/Hz and are phase locked to the internal 20 MHz TCXO. The plots in ***Figures 1*** and ***2*** show the measured phase noise of sample oscillators at 6.7 and 13 GHz, respectively, at offsets from the carrier between 100 Hz and 1 MHz.



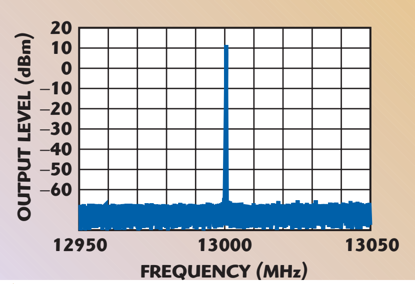
▲ *Fig. 1 Phase noise of a 6.7 GHz phase-locked oscillator.*



▲ *Fig. 2 Phase noise of a 13 GHz phase-locked oscillator.*



▲ *Fig. 3 Spur plot of a 6.7 GHz phase-locked oscillator.*



▲ *Fig. 4 Spur plot of a 13 GHz phase-locked oscillator.*

The modules are ideally suited for incorporation into modular designs for both commercial and military applications in the 5 to 14.55 GHz frequency range, particularly where size and space is at a premium. They are also suitable to be designed into special-to-type test fixtures, again where space is limited. Small, portable, battery-powered test sets can be constructed where a simple go, no-go test is required (on an aircraft radar system, for example).

The oscillator modules are fitted with a six-pin connector for providing DC power and lock alarm monitoring. They are supplied with a mating connector that allows them to be easily connected into existing or custom designs. In addition, +12 V DC operation at < 300 mA fits with the available power requirements of many OEMs.

APPLICATIONS

With a footprint of just 88 × 33 mm, the synthesized oscillators are small and versatile. They are suitable for applications such as the internal local oscillator for a block down-converter or up-converter for satellite monitoring or test systems where low phase noise and adequate mixer drive level are required. The synthesized oscillators can also be incorporated into existing receivers to act as independent frequency markers.

Very small size block converters can be constructed at an economic price by combining an ALO synthesized oscillator with a connectorised, double-balanced mixer and miniature filters. Their small size also makes the series ideal as the second, fixed local oscillator of an up/down conversion receiver design. Furthermore, the 5 to 6 GHz versions of the oscillators have sufficient output power at greater than +14 dBm to provide the common local oscillator for a pair of mixers in a direct conversion IQ receiver.

The oscillators can also form two-tone or multi-tone test sources that may be configured with combiners, filters and attenuators in suitable package formats for use in receiver or amplifier intermodulation distortion (IMD) test configurations. The good spectral purity, with regards to both spurs and harmonics, allow the modules to be configured without the need for external filters, although for some critical applications a low pass, harmonic filter may be desirable.

The plots shown in ***Figures 3*** and ***4*** are the measured spectra from two sample synthesized oscillators centered on 6.7 and 13 GHz, respectively. The span of ±50 MHz either side of the programmed centre frequency shows the absence of reference related spurs against a specification of < –60 dBc.

Inexpensive test sources for radar receiver or radar warning receiver (RWR) testing can be easily put together by selecting a suitable oscillator frequency and routing the output through a proprietary pulse modulator and step attenuator. In conjunction with a simple arbitrary waveform derived pulse generator, this would offer both minimum discernable sensitivity (MDS) and range tests in a single low cost test set. Similarly, stable, compact sources for TWT testing and aging can be constructed or provided as off-the-shelf bench test sets.

Additionally, more than one module combined with a multi-way switch would provide an easy-to-use test set for testing several sub-bands within a small volume package. For those applications requiring more output power, such as a short-range radiating test set, a range of amplifiers is available that can provide power levels of up to +30 dBm.

For higher frequency applications (above 14.5 GHz) the modules have sufficient output power to drive most frequency multipliers, which would allow operation up to at least 40 GHz or beyond, dependant upon connectors, and AtlanTecRF can supply a suitable range of amplifiers and frequency multipliers.

CONCLUSION

The new ALO series of synthesized local oscillator modules offers compact size and light weight, allied to modularity and a range of outputs that provide flexibility. These features make it a cost-effective and efficient option for many commercial and military applications, particularly where space is at a premium.

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QMA CONNECTORS WITH IMPROVED FREQUENCY RESPONSE, DURABILITY AND VARIETY

Two years ago a major US maker of two-way radios approached Times with this request: "Can you produce a cost-effective, fast mating, durable, long-life coax connector that offers steady RF performance if a mated pair is physically stressed?" The application is a high volume radio production final test. It is a controlled environment.

Objectives such as these are common across many users and markets that deal with RF communications. Through the years many schemes have been tried, but the attributes sought were not easily embodied in one product. "Fast mating" usually meant compromising on RF stability. "Durable and long life" meant being cost effective was unlikely. Trade-offs on one or more of the desired traits was inevitable, until today.

Times Microwave took a fresh look at the request, reasoning a solution would have broad appeal. The strategy was to improve upon an existing mating interface to save time and cost, and because the industry has accepted it. Several coax connector series were considered, but all failed at least one major hurdle until the QMA. The QMA plug incorporates the push-on, pull-off coupling action one associates more closely with pneumatics. This makes operation intuitive and easily mastered. The potential for cross threading, the need for wrenches and torque specs are now a thing of the past.

This user was concerned with frequencies below 2 GHz. While the industry standard for QMAs is 6 GHz, its size makes it a good candidate for improved frequency response and thus potentially attractive to users whose products operate at higher frequencies.

Size also plays a role in durability and cost. BNCs (the current choice) are of poor quality, not durable and vary electrically when stressed. The QMA is a good compromise between the MCX series, which is too small and delicate for production test, and the much larger, heavier and more costly to machine Type N. Most coax connectors ultimately need to attach to coax cable. The QMA's size keeps material costs, machining time and attachment design complexity for most popular coax cables within reason.

IMPROVEMENTS THAT MATTER

The first step was to improve durability to achieve a longer mating life for the QMA plug. The industry standard is 100 to 500 mates for OEM applications, beyond which (depending

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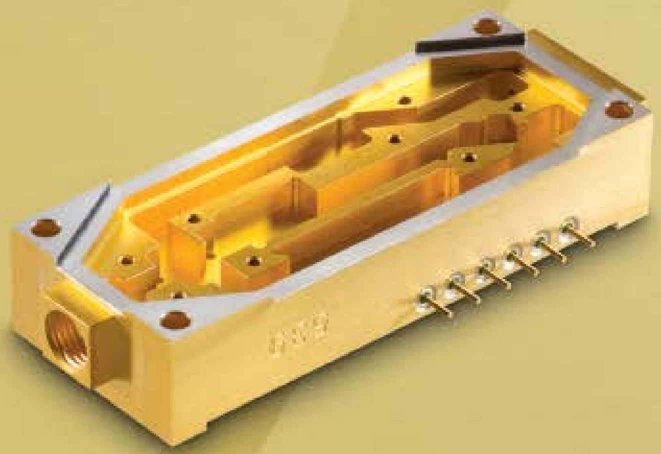

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on manufacturer) the retention force of a mated pair is reduced to unacceptable levels. In this production test application 500 mates would be reached in a few days. Therefore, the smaller spring fingers or tines within the male were replaced with fewer, but larger heat-treated tines. A special plating and polishing operation results in a smooth coupling nut action that lasts for 5000 mates when properly used. ***Figure 1*** shows the un-plated, industry standard tines on the left, and the un-plated Times Microwave Systems new design on the right.

Tolerances were tightened in several areas within the interface and compensation dimensions changed to increase the frequency response to 18 GHz and still achieve a return loss better than 18 dB in the straight configuration and approximately 17 dB in right angle configuration (Type N—QMA plug equipped cable assembly plus a QMA jack—SMA plug adaptor mated together).

▲ *Fig. 1 The unplated industry-standard tines (right) and the unplated Times new design (left).*

The new tine design requires more push-on force. There is a very distinctive "snap" when Times QMAs are mated. However, this solved another issue: RF measurement variations from lateral force or side load were eliminated or significantly reduced because a mated pair is held so tightly together. Well in excess of 20 lbs of pull force are required to pull a new pair apart if the coupling nut is not used to spread the tines before de-mating.

▲ *Fig. 2 The Times Microwave Systems' SureGrip coupling nut and its soft ergonomic ring.*

However, an unavoidable result was that a mated pair became very difficult to un-mate. To solve this problem, Times Microwave Systems developed its SureGrip™ coupling nut (see ***Figure 2***). Instead of smooth ridges it uses a sharp knurl. More importantly it employs a soft ergonomic ring. Together with the knurl it makes the coupling nut much easier to grip for retraction. Un-mating is now an easy task. The ring is quickly removed for tight, on-center applications, but a large machined ridge still exists that can be used to actuate the coupling nut.

Times Microwave Systems now offers both tine designs. Depending on the tine design, retentive force has been increased to over 40 lbs. Both carry the 5000 mate life cycle guarantee. Times QMAs will mate and un-mate 5000 times, remain within electrical specifications as stated on the data sheet and remain within a minimum of 2 lbs of retention force provided they are operated in accordance with recommendations. Should the need arise, Times designed a unique release tool to aid un-mating. It is compact to fit in

tight areas. The double end fits various Times QMA adaptors.

Since this was a high volume production test application, Times Microwave Systems fitted the new, more robust QMA plug to its highly popular SilverLine series of RF and microwave test cables. SilverLine Test Cables are highly durable, very stable with flexure and cost effective. This solved the issue of connectivity to radio test equipment on the production line.

▲ *Fig. 3 A sample of the SilverLine-QMA test cable and adaptor systems.*

▲ *Fig. 4 Times Microwave Systems' part no. 3191-197EA adaptor for making virtually any connector combination.*

Users who are considering adopting another connector series are also concerned with how to adapt to and from the new series without suffering significant RF degradation. The ability to accommodate a multitude of coax connector applications and even price and availability clearly enter into the decision-making process. To address these issues, Times Microwave Systems designed and built 28 unique QMA jack adaptors for the two-way, RF and microwave markets. This includes but is not limited to QMA jack to UHF and mini-UHF, BNC, Type N, SMA, TNC and others in both male and female. To eliminate that all-too-familiar performance degradation that comes from plating wear issues, all adaptors are US-made stainless steel. ***Figure 3*** displays a sample selection of Times Microwave Systems' new SilverLine-QMA test cable and adaptor systems.

Popular reverse polarity adaptors for wireless Internet and Cisco router applications also exist. Several between series adaptors outfitted with QMA plugs were designed and built that are useful for changing the ports on test equipment to a fast mating system, or simply to protect the original port connector from damage. Having tried and proven Times's QMAs, some popular test equipment manufacturers are now considering the QMA jack as an option that can be ordered for the RF port to eliminate damage from applying excessive torque, a main reason test equipment is returned for repair.

Two within-series adaptors have been developed that allow the user to create literally dozens or perhaps hundreds of coax series combinations or to connect multiple cables together, all through the QMA interface system (see ***Figure 4***).

Times now offers a two-watt QMA termination for amplifier and other test situations. To hold and protect all the new, high performance QMA adaptors, tools and termination, Times sells both a soft pouch and a hard case.

Times Microwave Systems, Wallingford, CT (203) 949-8400, www.timesmicrowave.com.

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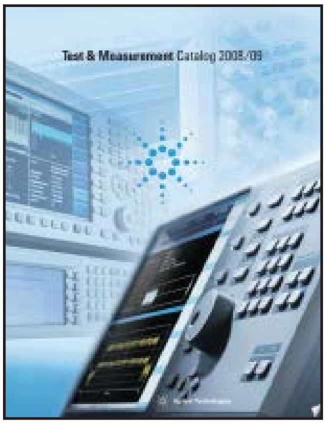

Test and Measurement Catalog

Agilent Technologies' Interactive Test & Measurement Catalog 2008/09 is now available online. The new interactive catalog has enhanced features, including the ability to send an e-mail to a colleague that links to a specific page within the catalog. Users can bookmark pages for ease of future reference. The entire catalog can be saved to a desktop or CD, enabling off-line viewing/portability. Also, the online catalog permits users to make personal reference notes on individual pages.

Agilent Technologies Inc.,
Santa Clara, CA (650) 752-5000, www.agilent.com.

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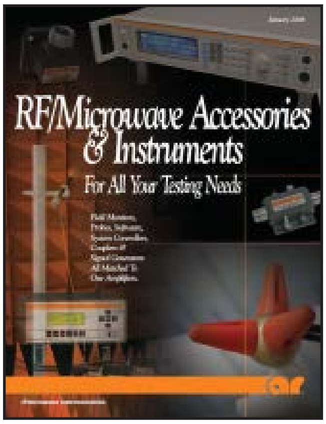

Accessories and Instruments Brochure

AR RF/Microwave Instrumentation's RF/Microwave Accessories & Instruments brochure highlights the company's signal generators, system controllers, directional couplers, EMC Test software, test cells and a complete line of field monitoring equipment. Also included is the full line of accessories for AR's conducted immunity test systems. Product photographs, descriptions and specifications are included for each model.

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

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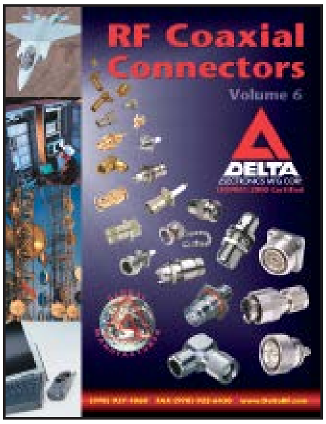

RF and Microwave Connectors

Volume 6 of Delta Electronics Manufacturing's full-line catalog contains over 200 pages of information on over 30 series of RF and microwave connectors, ranging in size from microminiature to large, high-power connectors. Along with common connector types, the catalog features many hard-to-find series, and a large selection of MIL-PRF-39012 QPL connectors and MIL-PRF-55339 qualified adapters.

Delta Electronics Manufacturing,
Beverly, MA (978) 927-1060, www.DeltaRf.com.

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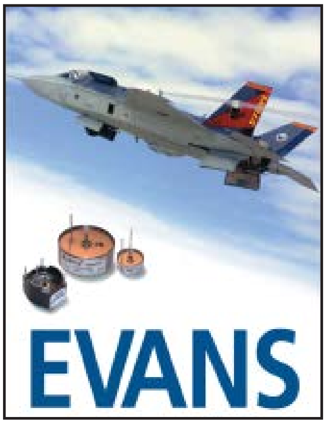

Online Capacitor Resource

A new Evans Capacitor Co. web site shares technical information, white papers, specs and pricing for its energy-dense hybrid capacitors and hybrid capacitor banks. High performance aircraft, including the Joint Strike Fighter and Apache helicopter, use Evans hybrid capacitors for laser targeting, communications modules, controls, cockpit displays, phased-array radars, fire control systems and more. The company's tantalum, hermetic, hybrid capacitors are over 4× the energy density of any military-style capacitor.

Evans Capacitor Co.,
East Providence, RI (401) 435-3555, www.evanscap.com.

RS No. 313

Designer's Guide

Hittite Microwave Corp. released its 13th edition Designer's Guide catalog for 2008. This publication includes 152 new digital, mixed-signal, RF, microwave and millimeter-wave product data sheets, as well as quality/reliability, application and packaging/layout information. Full specifications are provided for 633 products across 16 product lines including: amplifiers, attenuators, data converters, frequency dividers/detectors, frequency multipliers, high speed digital logic, mixers, demodulators/modulators, passives, phase shifters, power detectors, sensors, switches, synthesizers, VCOs/PLOs and VGAs.

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

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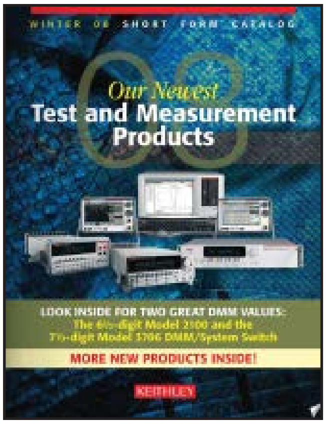

Online Catalog

Keithley Instruments Inc. announces the release of its 2008 Test and Measurement Product Guide. A short-form version of the guide is available online at www.nxtbook. com/nxtbooks/keithley/generalcatalog08/. This handy product guide offers details and specifications on Keithley's general-purpose and sensitive sourcing and measurement products, DC switching, RF switching and measurement, data acquisition solutions, semiconductor test systems and optoelectronics test hardware. Tutorials simplify choosing solutions for specific applications.

Keithley Instruments Inc.,
Cleveland, OH (440) 248-0400, www.keithley.com.

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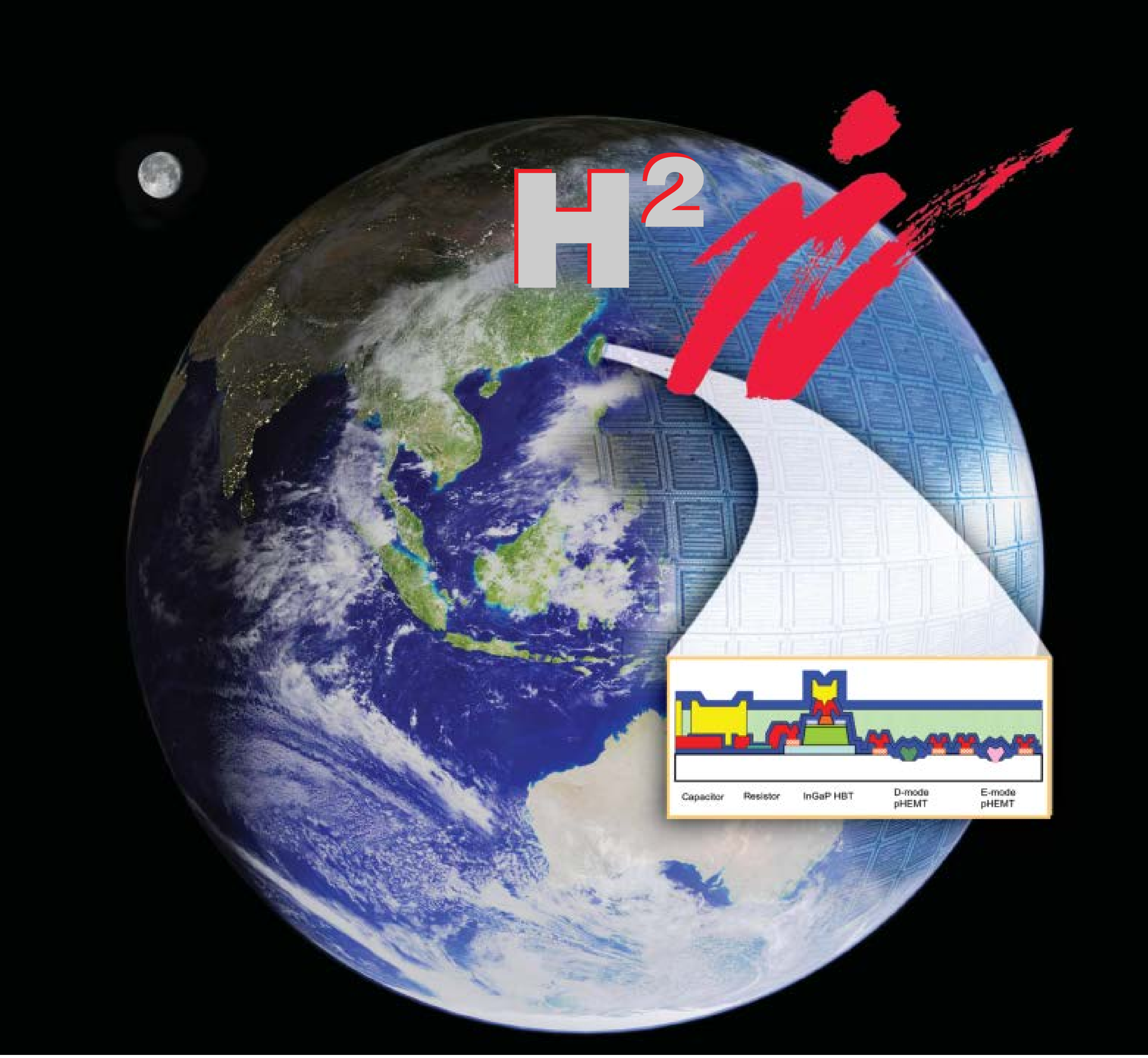
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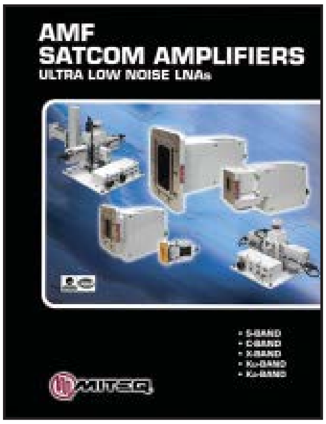
D-mode pHEMT
E-mode pHEMT



SATCOM Amplifier Catalog

MITEQ's AMF catalog line of SATCOM waveguide amplifiers offer the lowest noise figures available in an uncooled LNA in the various frequency bands associated with satellite communications. Achieving noise temperatures as low as 30 K for C- and S-band, 45 K for X-band, 60 K for Ku-band, and as low as 100 K for Ka-band, these amplifiers have been designed using state-of-the-art technology and can be used in either fixed or transportable applications.

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

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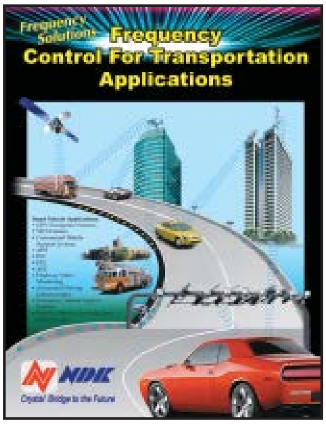

Frequency Control Devices

Nihon Dempa Kogyo Co. Ltd. (NDK), the supplier of frequency control devices to the transportation industry, has developed a new four-page color brochure highlighting its frequency control products for these applications. The brochure reviews in-vehicle and telematic applications. NDK's transportation components have proven to deliver excellent performance in the harshest of applications. Products covered in the brochure include crystals, TCXOs and XOs.

NDK America Inc.,
Belvidere, IL (815) 544-7900, www.ndkxtal.com.

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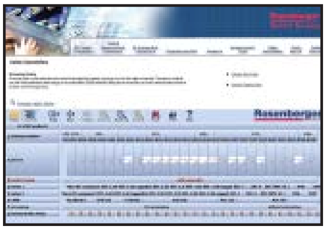

Cable Assemblies Brochure

A new Rosenberger tri-fold brochure features updates on the company's popular RFlex cable assemblies (standard connector configurations with ruggedized molded strain reliefs) and SMA+ (SMA connectors modified to operate to 36 GHz) cable assemblies. Also included are new product line additions including SMA Quick-Lock, SMA Push-On and RTK-028 cable assemblies. The SMA Quick-Lock provides a unique locking mechanism to replace threaded SMA male connections. The SMA Push-On offers a faster slide on connection. Both work with any SMA female connector. RTK-028 is Rosenberger's new triple shielded cable comparable to RG 316 DS but extremely flexible.

Rosenberger of North America LLC,
Lancaster, PA (717) 290-8000, www.rosenbergerna.com.

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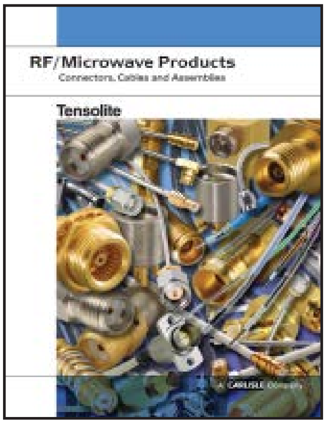

RF/Microwave Products

Tensolite's Connectors, Cables and Assemblies catalog is a reliable resource for high performance interconnect products. Renowned for quality and service, the products in this catalog display the wide range available to choose from for next generation systems. Vertical integration and extensive manufacturing capabilities ensure that quality and on-time delivery are never compromised.

Tensolite Co.,
St. Augustine, FL (800) 458-9960, www.tensolite.com.

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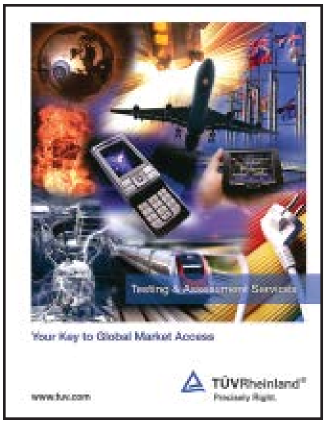

Testing and Assessment Services

TÜVRheinland has released a Testing & Assessment Services catalog for manufacturers planning to launch their products across the globe. The catalog details the company's testing services (e.g., EMC and Wi-Fi), product types (e.g., fuel cells and telecom), product approval schemes (e.g., Bluetooth and CB Scheme), service approval schemes (e.g., AS9100 and ISO 13485) and available certification marks (e.g., SEMI S2 and CE).

TÜVRheinland of North America Inc.,
Newtown, CT (203) 426-0888, www.us.tuv.com.

RS No. 320

RF/Microwave Components

Based in Korea and founded in 1999, Wave-Tech designs, manufacturers and delivers state-of-the-art, innovative RF signal processing components and sub-systems covering the frequency range of DC to 40 GHz. Wave-Tech's products include combiner, duplexer, BPF, BRF, coupler, splitter and filter-based assemblies for WiMAX, WiBro, GSM, DCS, PCS, CDMA, radar systems and wireless communications, and offer both custom and off-the shelf designs. Contact the company for a full-color brochure.

Wave-Tech Inc.,
Kyunggi-do, Korea +82 31 451 8774, www.wave-tech.co.kr.

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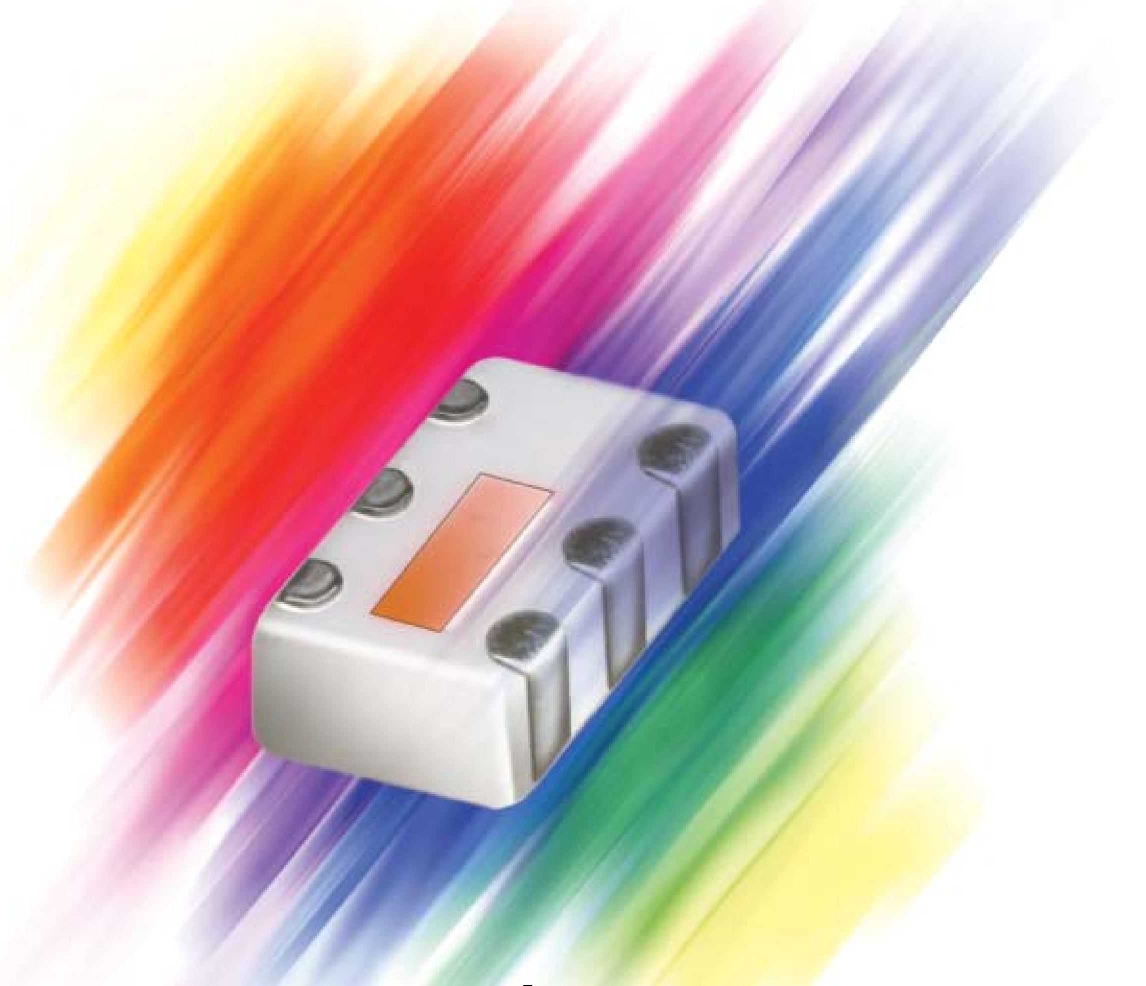
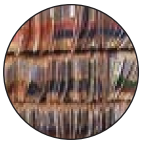
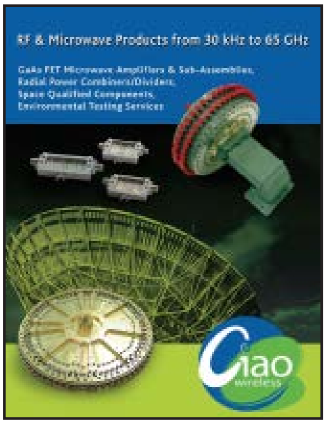




Catalog Update



Product Catalog

This catalog features the company's RF and microwave products that operate in a frequency range from 30 kHz to 65 GHz. Products include GaAs FET microwave amplifiers and subassemblies, radial power combiners/dividers, space qualified components and environmental testing services. All of the company's operations are located within a single modern and fully equipped 42,000 square foot facility.

Ciao Wireless Inc.,
Camarillo, CA (805) 389-3224, www.ciaowireless.com.

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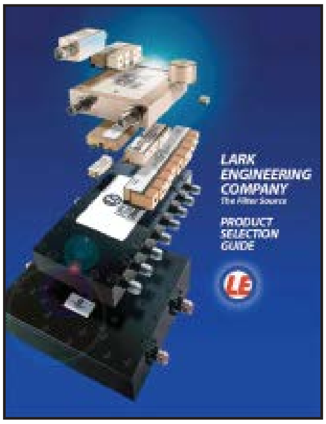

Filter Selection Guide

This product selection guide highlights the company's wide range of filter products, including its new switch filter systems. An eight-page short form catalog features a user friendly, quick reference to filter specifications and capabilities that guides users to the filter ideally suited for the application. For specifications and performance simulations, users are directed to a filter design tool located on Lark's web site.

Lark Engineering,
San Juan Capistrano, CA (949) 240-1233,
www.larkengineering.com.

RS No. 324

Signal Processing Components

This 144-page catalog offers comprehensive listings of RF, IF and microwave components with essential performance specifications for each product. In addition to the extensive component data, the catalog also provides a listing of Mini-Circuits' patents and the product model numbers to which they apply, as well as a complete listing of the National Stock Number (NSN) Guide for Mini-Circuits' considerable component collection.

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

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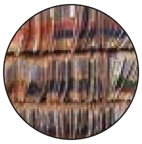
TRL CALIBRATION KITS • AUTOMATIC TUNERS • MANUAL TUNERS • TEST FIXTURES • ON-WAFER SOLUTIONS



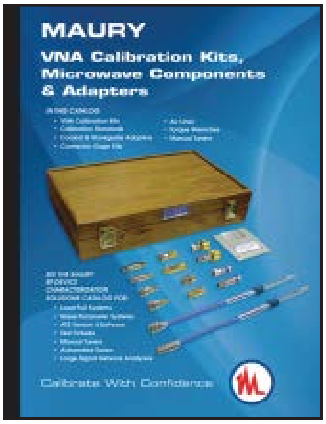
Catalog Update

RF Products Catalog

The E-store is an online guide to the company's most popular RF products. Product categories include fixed attenuators, terminations, manual variable attenuators, RF coaxial switches, power dividers/combiners and impedance matching. Users can search for products by keyword, name, or by browsing the illustrated, hyperlinked web pages.

JFW Industries Inc.,
Indianapolis, IN (317) 887-1340, www.jfwindustries.com.

RS No. 323



Components Catalog

This 146-page catalog features the company's entire line of precision coaxial and waveguide VNA calibration kits, microwave components and adapters, manual tuners, torque wrenches and connector gage kits. Entries feature detailed specifications, dimensional drawings, photographs, flange diagrams, ordering information and key reference materials. The catalog provides everything engineers need to find the best components for their applications. An interactive version of the catalog is available at the company's web site.

Maury Microwave Corp.,
Ontario, CA (909) 987-4715, www.maurymw.com.

RS No. 325

Semi-rigid Cable Products

This new data sheet features the company's Maximizer line of high performance semi-rigid cable products, available as custom assemblies, straight lengths or coils. Options include phase stable, low loss Maximizer Gold products and low loss, RG replacement Maximizer Silver products. The data sheet includes mechanical and electrical product specifications; IL, phase vs. temperature and power handling graphs; and design considerations.

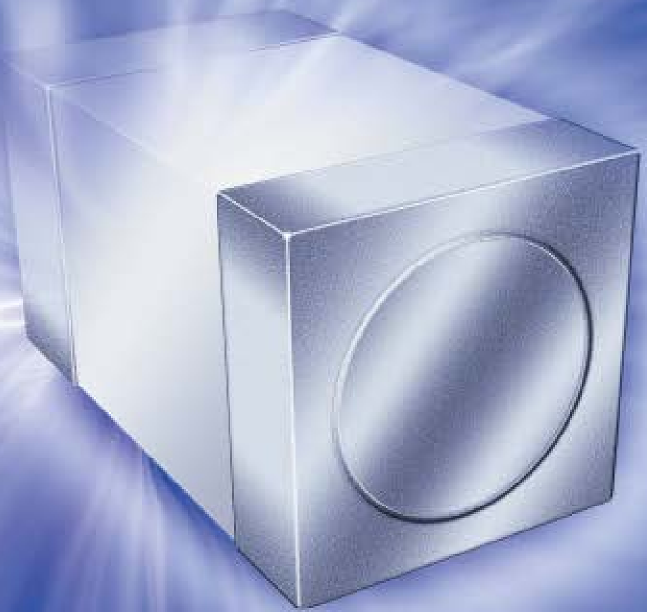
Teledyne Storm Products,
Woodridge, IL (630) 754-3361, www.stormproducts.com.

RS No. 327

One Powerful Diode from One Dynamic Company

Part Number	Voltage Rating Vb min, I = 10 uA Volts min	Total Capacitance V= 100 V, F = 1 MHz pF max	Carrier Lifetime I = 10 mA usec min	Series Resistance Rs @ 100 mA, F = 100MHz Ohms max
MMP7070	100	2.2	6	0.5
MMP7072	100	0.7	3	0.8
MMP7073	100	1.0	2.5	0.5
MMP7074	200	2.2	6	0.5
MMP7076	200	0.7	3	0.8
MMP7077	200	1.0	2.5	0.5
MMP7078	400	1.0	2.5	0.5
MMP7079	600	2.2	6	0.5
MMP7080	600	0.7	3	0.8

Over 20 models available. Call 603-641-3800 for details.



Teamwork. It turns great players into high charged service providers. And great products into an arsenal of integrated solutions. New high power MELF PIN diodes from Aeroflex's microwave products team, which includes Metelics and now MicroMetrics, are produced with a propriety glassing process. This creates large, full-face bonding surfaces on the anode and cathode, delivers low electrical and thermal resistance, and allows for power handling up to 100 Watts.

Models are available with:

- Voltage breakdowns up to 1,000 Volts
- Typical lifetime speeds of 1.0 to 8.0 µsec
- Resistances at 100 mA of 0.5 to 2.0 ohms

They're just one component in a dynamic line up. To see more, visit our Web site.

aeroflex.com/microwave

AEROFLEX
A passion for performance.

NEW WAVES: Amplifiers and Oscillators

Phase-locked Crystal Oscillator

The PLXO-Series phase-locked crystal oscillators from EM Research are now available with

significantly improved performance and smaller size. With the option of a surface-mount or connectorized package, PLXO offers several outputs: sine 7 dBm, 10 dBm, CMOS and PECL. The new smaller-sized surface-mount package (0.9 × 0.9 × 0.25 in.) also features exceptionally low phase noise characteristics (< –140 dBc/Hz @ 1 kHz typical) and enhanced spurious response (< –65 dBc). PLXO is available in fixed frequencies between 5 and 420 MHz with a lock range of ±5 ppm. Applications include DDS clocks, DAC and ADC clocks, fiber-channel ethernet, SAS/SATA, SONET/SDH and receiver LOs.

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

RS No. 216

Voltage-controlled Oscillator

Crystek's CVCO55CL-0902-0928 VCO operates from 902 to 928 MHz with a control voltage range of 0.5~4.5 V. This VCO features a typical phase noise of –115 dBc/Hz at 10 kHz offset and has excellent linearity. The model CVCO55CL-0902-0928 is packaged in the industry standard 0.5 × 0.5 in. SMD package. Input voltage is 5.0 V, with a max current consumption of 20 mA. Pulling and Pushing are minimized to 2.00 MHz and 0.80 MHz/V, respectively. Second harmonic suppression is –15 dBc typical. The CVCO55CL-0902-0928 is ideal for use in applications such as digital radio equipment, fixed wireless access, satellite communications systems and base stations.

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

RS No. 217

Ruggedized Rubidium Oscillator

This ruggedized rubidium oscillator is designed for ground tactical, shipboard and airborne applications where superior frequency stability under diverse environmental conditions is required. The 8200 supports these applications with superior phase noise and excellent short- and long-term frequency stability. Available as a small, low profile package, the 8200 Rubidium Oscillator provides 5 or 10 MHz output with multiple, high performance, low phase noise outputs support varied time and frequency applications; 1PPS input that enables easy integration with GPS receivers for improved accuracy; and a hermetically sealed package with shock/vibration hardening that enables operation over wide range of environmental conditions. Optional features include low phase noise, low g-sensitivity and multiple output configurations.

Symmetricom,
Santa Rosa, CA (707) 528-1230,
www.symmttm.com.

RS No. 218

Coaxial-resonator Oscillator

Z-Communications Inc. introduced its new CRO2935B-LF lead free, RoHS compliant, coaxial resonator oscillator in S-band (2850 to 3020 MHz) featuring low phase noise performance of –109 dBc/Hz at 10 kHz offset. This design offers typical tuning sensitivity of 23 MHz/V and is designed to provide 0 dBm output power (typical) at 8 V (DC) while drawing 26 mA (typical) over the extended operating temperature range of –40° to +85°C. The CRO2935B-LF comes in Z-Comm's industry standard MINI package measuring 0.50 × 0.50 in. × 0.22 in. This VCO is intended for automated surface-mount assembly and reflow. It is available in tape and reel packaging for production requirements.

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

RS No. 219

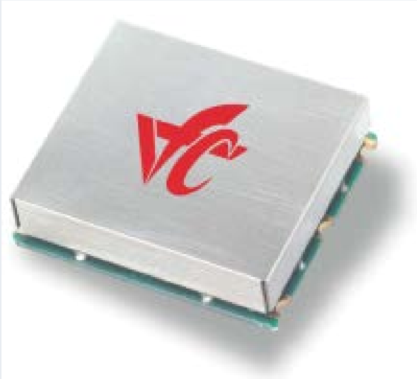
10 W Fixed Attenuators

Trilithic's RF & Microwave Components division released new fixed attenuators designed for 10 W DC to 6 GHz applications. This new fixed attenuator series is available in 3, 6, 10, 20 and 30 dB attenuation values. The Model HFP-510-6 features rugged construction with N male/female connectors, a maximum VSWR of 1.20:1 and attenuation accuracy of ±0.3 dB for values up to 20 dB. Applications include radar configuration, amplifier test and telecommunication and Wi-MAX labs.

Trilithic,
Indianapolis, IN (317) 423-6615,
www.trilithic.com.

RS No. 220

Jitter Attenuator



Valpey Fisher Corp. recently introduced the VFJA905 jitter attenuator, an integrated clock/PLL timing solution for 1 and 10 G synchronous ethernet applications. It features two LVCMOS outputs with a frequency of 25 MHz that can be locked to an input reference frequency. Two select inputs, S1 and S0, allow the user to select one of three preset input frequencies or a free-run mode. The VFJA905 provides a selectable input frequency range from 8 kHz to 200 MHz and an output frequency range of 10 to 200 MHz. It features ultra-low output jitter (0.18 ps RMS 12 kHz to 20 MHz). The device operates from a +3.3 V DC power supply and typically consumes 150 mW. The VFJA905 is available in a 19.5 × 15.5 mm surface-mount package and is RoHS 6/6 compliant.

Valpey Fisher Corp.,
Hopkinton, MA (800) 982-5737,
www.valpeyfisher.com.

RS No. 221

AIN Attenuator

Florida RF Labs announced the release of the 83A7038F surface-mount technology (SMT) attenuator. This new chip attenuator is especially suited for RF and microwave amplifiers and sub-systems. The 83A7038F SMT attenuator operates from DC to 6 GHz and provides excellent impedance match in the entire frequency band. The attenuator is rated at 20 W of continuous RF input power and it comes in a 0.200 × 0.175 in. (5.08 × 4.45 mm) package. It is available from 1 to 10 dB in 1 dB increments with a tolerance as low as ±0.3 dB.

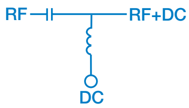
Florida RF Labs,
Stuart, FL (772) 286-9300,
www.rflabs.com.

RS No. 222

Detector Log Amplifier

Advanced Microwave is introducing ERD-LAR520E, a small wideband extended range

detector log amp for radar receiver systems. The unit is only 0.745 sq in. and covers 0.5 to 20 GHz with a wide dynamic range (72 dB). Other features include switchabili-



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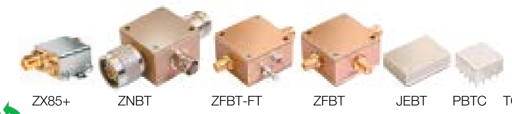
Mini-Circuits is your complete source for Bias-Tees, covering from 100 kHz to 12 GHz and handling up to 500mA DC in a variety of coaxial, plug-in, and surface mount packages. All of our Bias-Tees boast low insertion loss and VSWR. Our patented TCBT LTCC ceramic designs are the smallest in the world and are ready for your projects where very low price, space limitation, and temperature stability are a must. Our ultra-wideband ZX85 Bias-Tees use our patented Unibody construction to give you small size and high repeatability. Whether your applications call for biasing amplifiers, laser diodes, or active antennas, DC blocking, DC return, satellite communications, test, or if you have custom requirements, just contact Mini-Circuits and let us fit your needs to a "TEE"!

TYPICAL SPECIFICATIONS

Model	Freq (MHz)	Insertion Loss (dB)	Isolation (dB)	VSWR (:1)	Price $ea. Qty.10
TCBT-2R5G	20-2500	0.35	44	1.10	6.95*
TCBT-6G	50-6000	0.7	28	1.20	9.95
TCBT: LTCC, Actual Size .15"x.15", U.S. Patent 7,012,486.					**Qty.1-9**
JEBT-4R2G	10-4200	0.6	40	1.10	39.95
JEBT-4R2GW	0.1-4200	0.6	40	1.10	59.95
PBTC-1G	10-1000	0.3	33	1.10	25.95
PBTC-3G	10-3000	0.3	30	1.13	35.95
PBTC-1GW	0.1-1000	0.3	33	1.10	35.95
PBTC-3GW	0.1-3000	0.3	30	1.13	46.95
ZFBT-4R2G	10-4200	0.6	40	1.13	59.95
ZFBT-6G	10-6000	0.6	40	1.13	79.95
ZFBT-4R2GW	0.1-4200	0.6	40	1.13	79.95
ZFBT-6GW	0.1-6000	0.6	40	1.13	89.95
ZFBT-4R2G-FT	10-4200	0.6	N/A	1.13	59.95
ZFBT-6G-FT	10-6000	0.6	N/A	1.13	79.95
ZFBT-4R2GW-FT	0.1-4200	0.6	N/A	1.13	79.95
ZFBT-6GW-FT	0.1-6000	0.6	N/A	1.13	89.95
ZNBT-60-1W	2.5-6000	0.6	45	1.10	82.95
ZX85-12G+	0.2-12000	0.6	N/A	1.20	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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RF/IF & MICROWAVE COMPONENTS 395 Rev G ty from pseudo AC coupled to DC coupled and vice versa using a TTL control pin, which is useful for noisy environments and anti-jamming purposes. The ERDLAR520E is built using a proven and tested design technique and can handle 30,000 G force, equivalent to dropping the part from 10,000 feet to the ground and surviving the impact.

Advanced Microwave Inc., Sunnyvale, CA (408) 739-4214, www.advmic.com.

RS No. 223

Ku- and Ka-band Power Amplifiers

Endwave Corp. has announced the release of two new power amplifiers that deliver power and linearity performance for Ku-Band VSAT networks and emerging broadband satellite communications at Ka-Band. The Ku-band module features 40 dB (min) gain, 0.5 dB (max) gain flatness, typical output power at P1dB of +37 dBm, output 3rd order intercept of +47 dBm (typ) and DC current of 2.6 A at +15 VDC. The Ka-band module features 27 dB (min) gain, 1.0 dB (max) gain flatness, typical output power at P1dB of +36 dBm, output 3rd order intercept of +43 dBm (typ) and DC current at +15 VDC of 5.6 A. The noise figures of these transmit power amplifiers are equally impressive, measuring 4.5 dB (max) at Ku-band and 6.0 dB (max) at Ka-band. All units have a maximum VSWR of 2.0:1, without the use of RF input or RF output isolators. These parts are also available in a bench-top 19 in. rack-mount configuration for instrumentation applications. In this form-factor, the rack-mount units operate off 120-240 VAC and include all the necessary AC-to-DC conversion, power line filtering and conditioning internal to the unit.

Endwave Defense & Security Products Division, San Jose, CA (408) 522-3100, www.endwave.com.

RS No. 224

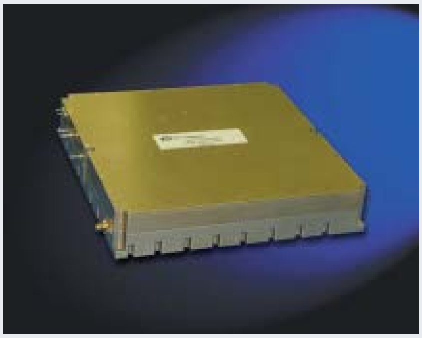
Low-noise Amplifiers

Avago Technologies announced a high-performance, low-noise amplifier (LNA) in an ultra-thin package size. Avago's new MGA-645T6 is an easy-to-use LNA with integrated bypass switch and shut down function optimized for customer premise equipment and base station equipment used in 802.11 b/g, WiMAX, WiBro and satellite digital multimedia broadband (S-DMB) applications. With a built-in bypass switch, Avago's MGA-645T6 operates in the 2.4 GHz spectrum; a follow-on product will support the 3.5 GHz spectrum. The MGA-645T6 LNA delivers an industry-leading low noise figure of 1.1 db and superior linearity. This improves the receiver's channel capacity and dynamic range and enables simultaneous WLAN and WiMAX functionality. Additionally, the LNA's shutdown function extends battery life of portable WiFi devices.

Avago Technologies, San Jose, CA (800) 235-0312, www.avagotech.com.

RS No. 226

GaN Broadband Power Amplifier



The Aethercomm Model Number SSPA 0.5-2.5-50 is a high-power, broadband, gallium nitride (GaN) RF amplifier that operates from 0.5 to 2.5 GHz. This PA is ideal for broadband military platforms as well as commercial applications because it is robust and offers high power over a multi-octave bandwidth. This amplifier operates with a base plate temperature of 85°C with no degradation in the MTBF for the GaN devices inside. It is packaged in a modular housing that is approximately 4.0 × 6.0 × 2.85 in. This amplifier has a typical P3dB of 50 watts at room temperature. Noise figure at room temperature is 6.0 dB typical. This amplifier offers a typical gain of 55 dB with a typical gain flatness of ±2.0 dB. Typical OIP3 is 55 dBm. Input and output VSWR is 2.0:1 maximum. Class AB current is ~4.0 amps typical employing a +24 VDC supply. This PA operates from 18 to 36 VDC input voltage. Typical harmonic values are –30 dBc in band.

Aethercomm Inc., San Marcos, CA (760) 598-4340, www.aethercomm.com.

RS No. 225

High Dynamic Range Driver Amplifier

M/A-COM, a business unit of Tyco Electronics, announced a new high dynamic range, low noise, single-stage driver amplifier. The MAAMSS0070 driver amplifier is intended for general-purpose applications in wireless infrastructure, such as mobile base station transceivers, cellular repeaters, WiMAX and WiBro customer premises equipment (CPE), and medical and test equipment. The usable bandwidth for this amplifier is 250 to 3800 MHz and is easily optimized for specific frequency bands through the use of external matching components. The MAAMSS0070 features output P1dB levels of +22 dBm with gain of 12 dB at 2140 MHz, while achieving high third-order output intercept performance and low noise figure over a broad frequency range. All the amplifiers in this family are housed in Pb-free, RoHS compliant SOT-89 plastic packages.

M/A-COM Inc., Lowell, MA (800) 366-2266, www.macom.com.

RS No. 227

High-linearity Amplifiers

Spectrum Microwave introduced a new line of lower cost amplifiers that provide high linearity and a low noise figure. These amplifiers feature a new lower cost surface-mount package and are available with frequencies from 20 to 4000 MHz, IP2 values to +75 dBm and noise figures as low as 0.9 dB. QBH-8900 series amplifiers offer gain from 10 to 28 dB and third-order intercepts to +45 dBm. The amplifiers feature internal blocking caps, biasing circuitry and RF matching. In addition to being cost effective, the new packaging for these high performance amplifiers is lightweight. QBH-8900 series amplifiers are designed for a wide range of applications including military radios, mobile communications platforms, point-to-point and multipoint radios, and military jammers.

Spectrum Microwave, Palm Bay, FL (888) 553-1531, www.spectrummicrowave.com.

RS No. 228

InGaP Power Amplifiers

RF Micro Devices announced the expansion of its wireless infrastructure product portfolio with the introduction of the SPA-1426Z and SPA-1526Z power amplifiers. The SPA-1426Z and SPA-1526Z are 1 and 2 W, respectively, indium gallium phosphide (InGaP) power amplifiers that address base station applications across all cellular standards and frequencies. The SPA-1426Z and SPA-1526Z Class A InGaP power amplifiers exhibit backed-off linearity performance, which is particularly critical to WCDMA high power amplifier applications. Housed in RFMD's proprietary SOF-26 package, the SPA-1426Z and SPA-1526Z provide low thermal resistance and power dissipation while providing RoHS and WEEE compliance. The SPA-1426Z and SPA-1526Z feature on-chip active bias circuitry and bias control pins, in addition to an input power detector and rugged class 1C ESD rating (greater than 1 kV HBM).

RF Micro Devices, Greensboro, NC (336) 664-1233, www.rfmd.com.

RS No. 229

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7/16
has hatched
A new composite 7/16
- 100% intermateable with standard 7/16
- 60% lighter
- Better Intermodulation withstanding
- No corrosion
- Reduction of your total cost of ownership
colors
RADIALL
The next conneXion
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COMPONENTS

MMIC SPDT Switch

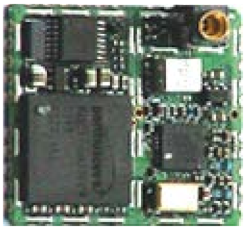
Mimix Broadband Inc. introduced a gallium arsenide (GaAs) monolithic microwave integrated circuit (MMIC) single-pole double-throw (SPDT) switch, the CSW0118-BD. Using 0.5 micron gate length GaAs pseudomorphic high electron mobility transistor (pHEMT) device model technology, this switch covers the 0.5 to 18 GHz frequency bands, and achieves 1.8 dB insertion loss and 35 dB isolation. The device also has a P1dB of 20 dBm and 2 ns rise/fall time. The CSW0118-BD is intended for radar, communications, avionics and test/measurement applications. The backside is gold plated, making the device compatible with either eutectic or conductive epoxy die attach and either thermocompression or thermosonic wire bonding. The chip also has surface passivation to protect and provide rugged parts with backside via holes and gold metallization to allow either a conductive epoxy or eutectic solder die attach process.

Mimix Broadband Inc., Houston, TX (281) 988-4600, www.mimixbroadband.com.

RS No. 232

Surface-mount GPS Module

The TiMax TI125 is a small OEM surface-mount GPS module that has been specifically designed for use in synchronization and timing applications. The TI125 has an on-board programmable NCO oscillator that outputs a synthesized frequency up to 30 MHz that is steered by the GPS receiver. The TiMax TI125 is an exceptionally small surface-mount package (25 × 27 × 4 mm) with a highly integrated architecture that requires the minimum of external components allowing easy integration into host systems.

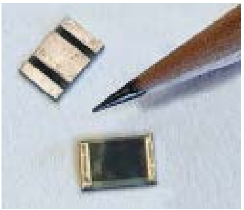


Connor-Winfield Corp., Aurora, IL (630) 851-4722, www.conwin.com.

RS No. 230

AIN Resistors

International Manufacturing Services Inc. (IMS) announced the release of the APX-CS series aluminum nitride resistors. These SMT devices feature a metal center stripe on the backplane to improve thermal conductivity through to the customer's groundplane, thus enhancing power capacity. The stripe is electrically isolated from either of the wrapped terminals. This design provides significant advantages to manufacturers of power supplies, two-way radios, industrial instruments, communications equipment and other systems that require high power dissipation, coupled with an environmentally safe solution. Enhanced performance can now be achieved without sacrificing real estate or employing a potentially hazardous substance.

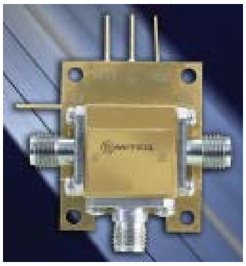


International Manufacturing Services Inc., Portsmouth, RI (401) 683-9700, www.ims-resistors.com.

RS No. 231

Digital SPDT Switch

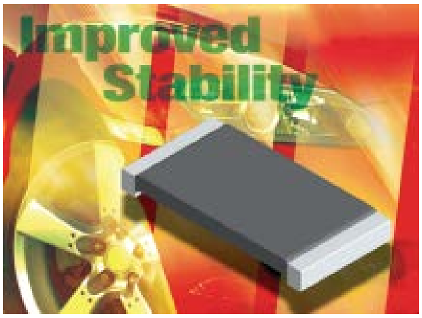
MITEQ Inc. introduced its new Model SW2-200305RD2NF reflective single-pole double-throw switch covering the frequency range from 20 to 30.5 GHz with a minimum of 30 dB isolation and VSWR of 2:1 (max). The 1 BIT input control word is TTL compatible and the time between the 50% point of the input control pulse to the 10 to 90 percent point of detected RF is within 30 ns. Rise/fall time is < 20 ns (typical). The power handling capability is 20 dBm. Available options include an extended band version, as well as sub-band versions.



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

RS No. 234

Surface-mount Resistor



Vishay Intertechnology Inc. announced a new surface-mount Power Metal Strip resistor that offers improved resistance stability during operation with a maximum resistance change of 0.5 percent through a 2000-hour workload. The WSLS2512 devices feature specially selected materials that allow for current sensing in high-temperature (+125°C), tightened-stability applications that require precision current monitoring of sensitive circuits, such as automotive electronic controls including engine, transmission and pollution controls. Manufactured using proprietary techniques that result in extremely low resistance values of 0.01 to 0.1 Ω, the WSLS2512 features all-welded construction with a solid metal nickel-chrome alloy resistive element. The resistor offers low inductance values of 0.5 to 2.0 nH, a tight tolerance down to ±0.5 percent and low TCR values down to ±30 ppm/°C.

Vishay Intertechnology Inc., Malvern, PA, www.vishay.com.

RS No. 236

SUBSYSTEMS

Desktop Electroplating System

LPKF Laser & Electronics introduces the MiniContac RS, reverse pulse plating system, specially developed for the professional production of prototype and small batch production PCBs. This cost-effective system is completely enclosed in a compact table top size, ideal for any rapid PCB prototyping situation, especially small runs and tight work locations, such as a research environment. The MiniContac RS has the ability to plate holes as small as 8 mil (0.2 mm) vias in 62 mil thick standard PCBs smoothly; thin or fragile materials can easily be placed in a support framework before processing. The MiniContac RS handles circuit boards as large as 9 × 13 in. (230 mm × 330 mm) and is completely closed with no external connection needed.

LPKF Laser & Electronics, Wilsonville, OR (503) 454-4212, www.lpkfusa.com.

RS No. 237

4.9 GHz Sector Antennas

Radio Waves Inc. has announced a family of new 4.9 GHz sector antennas that are optimized for the 4.9 to 5.0 GHz band. These antennas, the SEC-49 series, are available in both 60 and 90 degrees and are available as vertically or horizontally polarized sectors. These new antennas cover a number of services in the public safety band. This new band will aid communications among public safety agencies by making communications easy for emergency workers in different jurisdictions.

Radio Waves, N. Billerica, MA (978) 459-8800, www.radiowavesinc.com.

RS No. 238

Remote Optical Network Units

Andrew Corp. has introduced two new remote units for cost-effective radio-over-fiber applications that ensure coverage and capacity in demanding wireless environments. The new ION-B High-power Single-band UMTS remote unit and ION-B Medium-power Triple Band EGSM/GSM1800/UMTS remote unit offer higher power options for extending in-building radio coverage and enhancing performance in multi-carrier and multi-band applications. Andrew's Intelligent Optical Network (ION) series optical distribution system carries radio signals loss-free over fiber optics at all power levels at distances ranging from 100 meters to 20 kilometers.

Andrew Corp., Westchester, IL (708) 236-6600, www.andrew.com.

RS No. 239

High-speed Fabric Interface

Emerson Network Power's new Embedded Computing business, formerly Artesyn Communication Products, announced a new high-speed switched fabric interface for its KAT4000 AdvancedTCA (ATCA) telecom blade. The enhanced interface provides dual 10-Gigabit Ethernet channels for accessing the ATCA high-speed fabric. The KAT4000 is a configurable, PICMG 3.1 (Option 9) ATCA blade that can accommodate up to four AdvancedMC modules. To maximize system throughput and flexibility, the KAT4000 provides separate control/management and data planes, each with its own independent switching capability and ATCA fabric connection. The control/management plane, available with an optional Freescale 8548 management processor complex, utilizes separate Gigabit Ethernet and PCI Express switches to connect each AdvancedMC site to the ATCA Base Fabric. The KAT4000's data plane utilizes Gigabit Ethernet (one channel per site) to connect each AdvancedMC site to an on-board switch, and dual 10-Gbit/sec Ethernet channels to link the switch with ATCA's 10-Gbits/sec high-speed fabric.

Emerson Network Power, Embedded Computing, Madison, MI (608) 831-5500, www.artesyncp.com.

RS No. 241


THE MILITARY AND AEROSPACE INDUSTRIES RELY ON ACC ADVANCED SWITCH TECHNOLOGY.
Our complete line of switch products include:
• PIN Diode and GaAs FET Design
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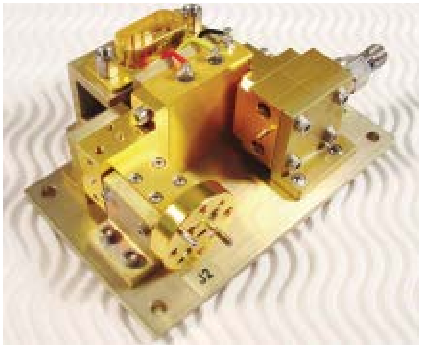
NEW PRODUCTS

W-band Multiplier

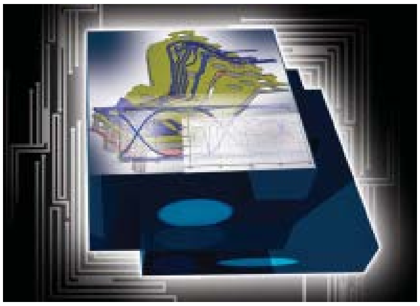


The Spacek Labs Model A3X2X91F W-band multiplier assembly is available space qualified. The A3X2X91F consists of a X2 frequency multiplier feeding a power amplifier followed by a X3 frequency multiplier. A combination high pass/low pass waveguide filter provides suppression of unwanted harmonics for this W-band multiplier. This LO source multiplier drove the Spacek Labs manufactured mixers for the transmitter and receiver front-end of the CloudSat Program Cloud Profiling Radar. Launched on April 28, 2006, and now orbiting at an altitude of 438 miles, CloudSat has been providing unprecedented, vertically cross-sectioned images of major weather systems.

Spacek Labs Inc.,
Santa Barbara, CA (805) 564-4404,
www.spaceklabs.com.

RS No. 242

Simulation and Design Software



Ansoft Corp. has released new versions of Nexxim,® the company's high-capacity circuit simulation software, and Ansoft Designer,® its integrated schematic and design management software. The products include new statistical analysis and transient simulation capabilities that allow engineers to rapidly simulate high-speed serial channel behavior. Nexxim v4 features VerifEye,® a new methodology for eye analysis of serial links using statistical methods that maintain accuracy while offering major reductions in run time compared to conventional transient methods. These products combine to allow engineers to accurately predict signal integrity effects and EMI/EMC performance of advanced electronic systems, including gigabit communication channels, multi-function high-speed wireless systems and sophisticated microwave systems.

Ansoft Corp.,
Pittsburgh, PA (412) 261-3200,
www.ansoft.com.

RS No. 243

Digital-pulse Compression Subsystem

Temex announced the launch of a new digital pulse compression subsystem (DPCSS) with chirp generator (transmitter) and compressor (receiver) modules at intermediate frequency (IF): the CI BM7xx. This DPCSS with analog I/Os enables upgrades of SAW-based radars by increasing overall system performance. The CI BM7xx comes from Temex's 30 years of experience in the design and manufacturing of high-performance signal processors. Combining the latest generation field-programmable gate arrays (FPGAs) for signal processing and the analog input and output for the IF of the SAW, this signal processor is an excellent replacement solution for currently operating SAW-based radars. Temex's digital solution lowers power consumption in comparison to oven devices. This pulse compression sub-system is self-consistent and needs only a DC supply voltage for operation. All the pulse compression parameters such as center frequency, bandwidth, time dispersion, modulation law and weighting function are programmable.

Temex,
Sophia-Antipolis Cedex, France,
+33 (0)4 97 23 32 53, www.temex.com.

RS No. 244

TEST EQUIPMENT

Lightweight Signal Generator

AR RF/Microwave Instrumentation has introduced a lightweight 2U, 3.5 in. high signal generator with a bandwidth of 9 kHz to 1.2 GHz and an RF output range from –140 to +13 dBm. The Model SG1200 provides very low carrier frequency with excellent resolution and low harmonics. The SG1200 is suited for RF immunity testing for both conducted and radiated immunity systems covering the requirements for MIL-STD-461D/E, DO-160D/E, IEC 61000-4-3, IEC 61000-4-6 and many other EMC test standards. The SG1200 signal generator features both internal and external AM, FM, phase and external pulse modulations, and provides the capability of combining internal and external modulations. The SG1200 is a rugged design providing electronic trip protection to protect the generator's output against reverse power up to 50 W.

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

RS No. 245

Ruggedized Synthesizers

Elcom Technologies Inc. has released its new RS Series family of ruggedized synthesizers. The single-loop PLL synthesizers employ Elcom's proprietary design architecture to provide low phase noise and fast switching speed, < 50 microseconds, in a compact ruggedized configuration. These synthesizers maintain specified performance under conditions of vibration exceeding 6 G-RMS. Operating base plate temperature range is –55° to +85°C. The PLL synthesizer module includes GaAs ICs assembled with chip-and-wire construction. For high reliability, the module is hermetically sealed with laser welded RF and DC connectors, and with laser welded top and bottom covers. The RS Series PLL synthesizers are intended for low observability in spread-spectrum military airborne and ground-based communications systems. They have application in many types of radar and ECM systems and in other types of communications systems.

Elcom Technologies Inc.,
Rockleigh, NJ (201) 767-8030,
www.elcom-tech.com.

RS No. 246

Broadband Vector Network Analyzer

Anritsu Corp. has enhanced its Lightning Broadband ME7808C vector network analyzer (VNA) so that it provides higher output power to drive active and passive components at optimum measurement levels, and improved calibration stability so that the broadband VNA can satisfy the most demanding measurement needs. Among the design enhancements made to the ME7808C are relocating the couplers inside the multiplex combiners, improved Kelvin design bias tees, better RF and LO stability, and greater short-term and long-term calibration and measurement stability. The ME7808C has also been designed with improved phase stable cables for better calibration and measurement stability. The ME7808C delivers significantly improved performance due to the enhancements. It can provide up to –13 dBm of power across the entire 40 MHz to 110 GHz frequency range.

Anritsu Co.,
Morgan Hill, CA (408) 778-2000,
www.anritsu.com.

RS No. 248

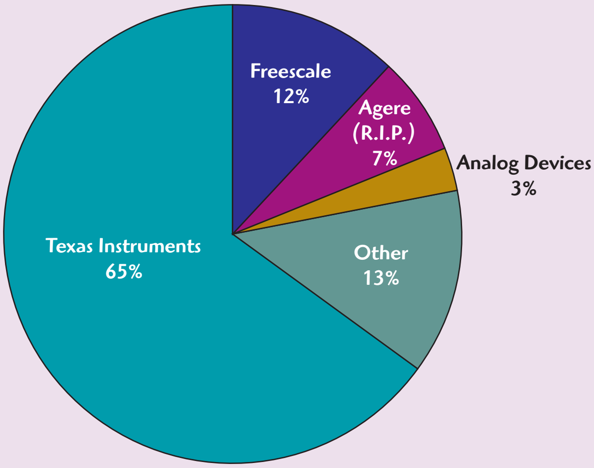
The International Microwave Symposium is the headline conference of the IEEE Microwave Theory and Techniques Society (MTT-S). This will be the largest technical Conference to be held in Atlanta in the next two years and will feature a large trade show as well as a wide variety of technical papers and workshops. The IEEE MTT-S International Microwave Symposium 2008 (IMS2008) will be held in Atlanta, GA, Sunday, June 15 through Friday, June 20, 2008, as the premiere event of Microwave Week 2008.

Microwave Week 2008: The IMS 2008 technical sessions will run from Tuesday through Thursday of Microwave Week. Workshops will be held on Sunday, Monday and Friday. In addition to IMS2008, a microwave exhibition, a historical exhibit and the RFIC Symposium (www.rfic2008.org) will also be held in Atlanta during Microwave Week 2008.

Microwave Metrics

DSP Vendor Market Shares

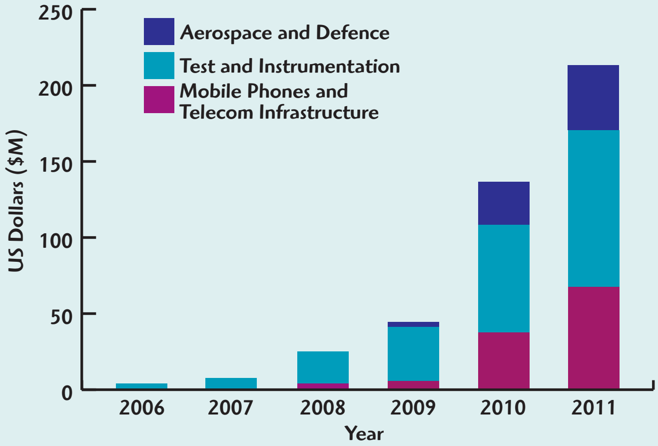
Total DSP chip shipments in 2007 were down by some 6.7% to the $7.8 B level. Actually, unit shipments were up 6.4%, but the ASP decline was 12%, dragging down the revenue figure.



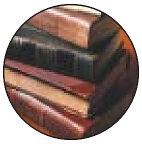
Source: Forward Concepts, 1575 W. University Drive, Suite 111, Tempe, AZ 85281-3283 (www.fwdconcepts.com)

RF MEMS Switch Market Forecast

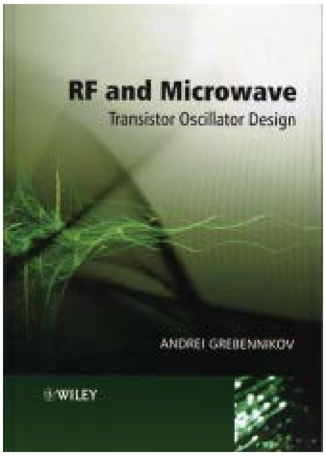
WTC (Wicht Technologie Consulting) says RF MEMS devices will quickly grow to $210 M in 2011. More than $100 M of this market is destined for test and instrumentation, e.g. automated test equipment used in the semiconductor industry, followed by devices used in mobile phones and telecom infrastructure. At the module level, reconfigurable power amplifiers and antenna modules for cell phones should exceed $150 M in 2011.



Source: Wicht Technologie Consulting, Frauenplatz 5, D-80331 Munich, Germany (www.wtc-consult.de)



RF and Microwave Transistor Oscillator Design



Andrei Grebennikov
456 pages; $150 • John Wiley & Sons Ltd.
ISBN: 978-0-470-02535-2

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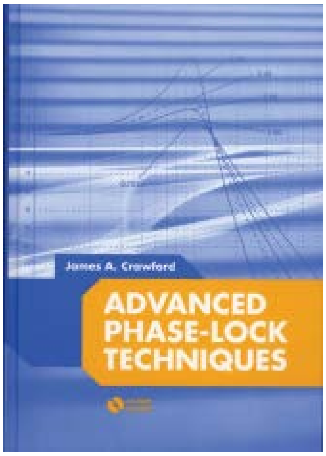
The main objective of this book is to present all relevant information necessary for RF and microwave transistor oscillator design, including well-known and new theoretical approaches and practical circuit schematics and designs. Chapter 1 presents the most commonly used design techniques for analyzing nonlinear circuits, in particular transistor oscillators. Chapter 2 introduces the principles of oscillator design, including start-up and steady-state operation conditions, basic oscillator configurations using lumped and transmission-line elements, and simplified equation-based analysis and design techniques. Several examples of stability criteria for different single-resonant and double-resonant oscillator circuits are described and analyzed in Chapter 3. Chapter 4 presents both the empirical and analytical optimum design approaches applied to series and parallel feedback oscillators, including circuit design and simulation aspects and high efficiency. Chapter 5 describes different oscillator noise models to express a clear relationship between the resonant circuit and active device noise model parameters. Chapter 6 discusses varactor modeling issues, varactor nonlinearity and its effect to frequency modulation, and resonant circuit techniques to improve VCO tuning linearity using lumped and transmission-line elements. Chapter 7 discusses the technological aspects to realize MOS varactors and spiral inductors, basic concepts of circuit design and implementation issues. Chapter 8 discusses the basic concepts of wideband VCO design and gives specific circuit solutions using lumped elements and transmission lines to improve their frequency tuning characteristics. Chapter 9 discusses phase noise reduction techniques and gives specific resonant circuit solution using lumped and distributed parameters for frequency stabilization and phase noise reduction.

Advanced Phase-Lock Techniques



James A. Crawford
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Phase-lock techniques in time and frequency control systems have expanded substantially since the author's first book *Frequency Synthesizer Design Handbook*, (Artech House) that was published in 1994. This text builds on the foundational material that was provided in the original text, with expanded attention given to a range of topics that are germane to phase-lock systems. Chapter 1 offers a broad-brush perspective of the phase-lock technique. A number of helpful formulas have been collected in Chapter 2 for easy and quick reference. Chapter 3 is motivated by the rather obvious need to always know what the theoretical limits are concerning a given design problem. In general, design problems that demand performance close to theoretical limits will require greater design precision, power consumption and/or development effort. Chapter 4 describes many different noise sources and their analysis techniques. The role that noise plays at the system level is considered in Chapter 5. Chapter 6 provides a discussion about classical continuous-time PLLs used for frequency synthesis, quickly moving into a presentation of pseudo-continuous systems that are more representative of modern systems. Chapter 7 is an expanded discussion of sampled-PLLs that should be firmly connecting the concepts between continuous-time and sampled-time PLLs in a rigorous manner. With the proliferation of the fractional-N frequency synthesis method and Δ-Σ methods in general, Chapter 8 opens with a historical development time-line for fractional-N followed by fractional-N frequency synthesis. Chapter 9 is devoted to a fairly extensive discussion about oscillators. Chapter 10 stems from a number of development projects pertaining to precision clock and data recovery (CDR) systems, also called bit synchronization.

Career Corner

The Seven Good Years and What's After

The boost in electromagnetic-related applications is shared across a number of industries, resulting in a growing demand for RF engineers. A shortage of RF engineers has been reported for growing industry sectors in both consumer and defense markets. RF engineering has been a fast growing segment since 2005. Stability and wages for RF engineers have improved in recent years and are projected to continue until 2012. The demand exhibits a distinct preference towards experience and hands-on RF design skills. This implies heavy RF/microwave R&D across the board. During these seven (good) years, companies and engineers are engaged laying foundations for future microwave technologies and infrastructures in various geographies and markets.

The diverse range of industries now sharing this RF trend simultaneously indicates the stability of this process. Apparently the world is "going wireless" this time (unlike the early 2000s) and with an even broader scope of applications. These seven years will reflect that. The projected duration of the growth and intense R&D activity allows (and maybe calls for) long-term planning. Awareness to the process will challenge forward-looking employers to develop resourceful plans to meet their long-term staffing needs. Career-minded RF engineers will make wise career decisions including the relation between areas of expertise, technologies and markets in their considerations.

This is also a call to the academy and education institutions to realize the potential as well as the urgency of the situation. Using the term **URGENT** as the adjective is not an understatement and is not related to corporations' need to compete and sustain market shares. It is about the jobs of the future. Beyond meeting technology goals, the course of the seven good years will result in a new order of technology superpowers. Corporations may be multinational. The employees are always local somewhere. Corporations establish R&D centers where the talent is, and RF engineering is becoming key technology for numerous future uses.

A society that will develop more and better engineers will provide its industry with a better selection of creative talent and will ensure its future technological leadership and the jobs that go along with it. *Note: Although technology can be sold or transferred, the creative resources behind it are not transferable (by the time the technology is sold they are already engaged inventing the next generation). This is the ultimate product of education.*

A consortium of electrical engineering faculties, industry leaders and related engineering associations would make a fine start. *Microwave Journal* and ElectroMagneticCareers.com encourage further discussion of this topic and will provide the stage for it. Readers' feedback is welcome.

Isaac Mendelson
ElectroMagneticCareers.com
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FAX: +49 7122 828145
bberanek@horizonhouse.com

Israel
Oreet Ben Yaacov
Oreet International Media
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obenyaacov@horizonhouse.com

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FAX: +82 2 481-3414
yschinn@horizonhouse.com

Japan
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Ace Media Service Inc.
12-6, 4-Chome,
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	Description	Device	Frequency (GHz)	Gain (dB)	Gain Flatness (dB)	Noise Figure (dB)	Output P1dB (dBm)	OIP3 (dBm)	Bias (mA @V)	Package (mm)
5-10 GHz Mimix SmartSet	Buffer Amplifier	XB1007-QT	4.0-11.0	23.0	+/-1.5	4.5	+20.0	+30.0	100 @ 4.0	3x3
	Power Amplifier	XP1035-QH	5.9-9.5	26.0	+/-1.0	-	+29.0	+39.0	500 @ 6.0	4x4
	Power Amplifier	XP1039-QJ	5.9-8.5	15.0	+/-1.0	-	+34.5	+49.0	1150 @ 8.0	6x6
	Receiver	XR1011-QH	4.5-10.5	13.0	+/-1.0	1.8	+6.0	+16.0	130 @ 4.0	4x4
	Doubler	XX1002-QH	5.0-12.0 fout	16.0	+/-1.5	-	+16.0 Psat	-	125 @ 5.0	4x4
	Transmitter	XU1012-QH	5.0-10.0	-8.0	+/-1.0	-	+7.0	+17.0	120 @ 4.0	4x4
10-16 GHz Mimix SmartSet	Buffer Amplifier	XB1008-QT	10.0-21.0	17.0	+/-2.0	4.5	+19.0	+32.0	100 @ 4.0	3x3
	Power Amplifier	XP1042-QT	12.0-16.0	21.0	+/-1.0	-	+25.0	+38.0	500 @ 5.0	3x3
	Power Amplifier	XP1043-QH	12.0-16.0	20.0	+/-1.0	-	+30.0	+41.0	700 @ 7.0	4x4
	Receiver	XR1007-QD	10.0-18.0	13.5	+/-1.0	2.7	+5.0	+15.0	150 @ 5.0	7x7
	Doubler	XX1000-QT	15.0-45.0 fout	10.0	+/-2.0	-	+18.0 Psat	-	200 @ 5.0	3x3
	Doubler	XX1002-QH	5.0-12.0 fout	16.0	+/-1.5	-	+16.0 Psat	-	125 @ 5.0	4x4
	Transmitter	XU1014-QH	8.0-18.0	-10.0	+/-1.0	-	+2.0	+12.0	80 @ 4.0	4x4

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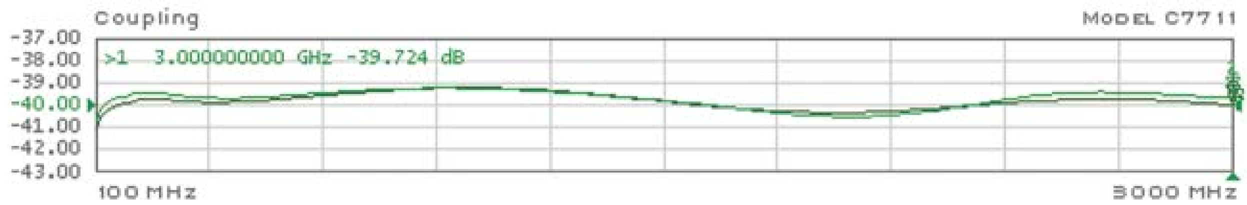

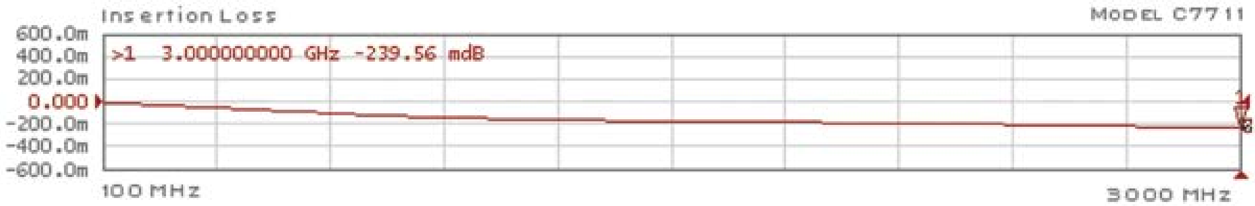

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Model	Coupler Type	Frequency (MHz)	Power CW (Watts)	Coupling (dB)	Flatness (±dB)	Insertion Loss (dB)	VSWR (Mainline)	Directivity (dB)	Size (Inches)
C7734	Dual Directional	30-2500	100	43	±1.5	0.35	1.25:1	18	3.5 x 2.6 x 0.7
C7148	Bi Directional	60-600	200	10	±1.0	0.35	1.20:1	20	6.0 x 4.0 x 0.75
C7711	Dual Directional	100-3000	100	40	±1.0	0.35	1.25:1	18	3.0 x 2.2 x 0.7
C7783	Bi Directional	200-1000	200	20	±0.75	0.2	1.20:1	20	3.0 x 1.5 x 0.53
C6600	Bi Directional	200-2000	200	20	±1.2	0.25	1.25:1	18	4.0 x 2.0 x 0.72
C7152	Bi Directional	300-3000	100	20	±1.0	0.35	1.20:1	15	3.7 x 2.0 x 0.75
C7811	Dual Directional	500-2500	100	40	±0.5	0.2	1.25:1	20	3.0 x 2.0 x 0.6
C7753	Bi Directional	700-4200	100	20	±1.0	0.35	1.25:1	18	1.8 x 1.0 x 0.6