Multiharmonic Manipulation for Highly Efficient Microwave Power Amplifiers

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Received 8 January 2001; accepted 24 April 2001

ABSTRACT: Multiharmonic manipulation is presented as the most effective solution to improve power amplifier (PA) efficiency performances. Remarkable improvements in output power, power gain and power-added efficiency (PAE) are demonstrated, properly manipulating the input and output second and third harmonics, as compared to more classical design approaches. Experimental results at 5GHz confirm the feasibility, the validity and effectiveness of the proposed approach, increasing the maximum measured power-added efficiency from 39% to 61%. © 2001 John Wiley & Sons, Inc. Int J RF and Microwave CAE 11: 366–384, 2001.

Keywords: power amplifiers; high efficiency; low voltage; high linearity; multiharmonic manipulation

I. INTRODUCTION

The power amplifier (PA) stage is the crucial element in many microwave systems, including handy phone applications, satellite payloads, microwave transponders and many others. The key specification of the power stage imposes minimum power consumption (to obtain long talk and standby operation), primarily in subsystems where the PA dissipated power is the major source of power consumption. High-efficiency performance becomes therefore the major challenge of the design; high efficiency in turn implies a higher output power for the same DC consumption or, conversely, a remarkable reduction of the DC power producing the same output power level. As a consequence, reduction of the size and weight of battery cells and heat sinks is ensured. Moreover, the reduced power dissipation guarantees a lower device junction temperature with a consequent improvement in reliability [1] of the stage.

The efficiency of microwave PAs is limited by the active device parameters and operating conditions. A major progress has been achieved on GaAs power FET device performances and fabrication technology, from the design side, it is well established that a high efficiency can be obtained by a proper selection of bias point and harmonic terminations or, from a different point of view, by a proper output voltage and/or current waveform shaping. In particular, maximum output voltage must occur at low (zero) current levels and maximum current must correspond to very low voltages, thus minimizing dissipated power on the active device. Starting from the pioneering work of Snider in 1967 [2] and as a natural evolution of classical and well-known A, AB, B and C biasing classes, many “harmonic terminating strategies” have been proposed in literature. Some of them, like the Class-F approach [2–6] or the more unusual Harmonic Reaction Amplifiers [7–9], focus on the output network terminations only. Alternatively, different approaches based on
device switching mode operation, originated new operating classes, as Classes D (or S) [6] and E [10–14], and demonstrated to be extremely effective for low operating frequencies (less than a few GHz).

In recent years, demanding applications at higher operating frequencies forced many designers to a critical review of the basic assumptions underlying the above-mentioned strategies. In fact, design approaches based on device switching-mode operation lose a major part of their effectiveness when designing high efficiency microwave or millimeter-wave PAs, mainly due to the difficulty in controlling more than a couple of voltage or current harmonic components beyond an output shorting capacitive behavior of the active device itself. On the other hand, other aspects like the influence of input harmonic terminations different from the normally used short-circuiting solution have to be considered. As an example, the second harmonic terminating control schemes, both at input and output ports, are fruitfully employed [15–20] or particular driving waveforms have been suggested [21]. Unfortunately, in open literature, such second harmonic approaches (mainly experimental and with a lack of physical insight) often resulted in confusing contributions, reporting even contradictory results and conclusions.

In this article, a comprehensive theory of a multiharmonic manipulation design strategy is attempted. Moving from a weighting procedure for the second- and third-order harmonic output voltage components, an effective improvement of microwave PA performances is demonstrated in terms of large-signal gain, output power and power-added efficiency (PAE).

Measurements performed on a sample PA, designed using the proposed technique to operate at 5 GHz and realized in hybrid form, will be presented and compared to the performances of an optimized companion PA, designed using a classical Tuned Load (TL) approach. The results clearly demonstrate both the feasibility and effectiveness of the proposed methodology.

II. BASIC ASSUMPTIONS

Every active device used as an amplifying element exhibits major power limiting mechanisms, leading to output power saturation; in field-effect based devices, such physical constraints reside in the gate–source junction forward conduction and channel pinch-off (determining the maximum current swing), together with triode region and gate–drain junction breakdown (fixing the maximum voltage swing).

A careful technological optimization of doping profiles and gate recess can alleviate the effects of the above limitations, but a further power performance improvement however has to be based on smart design methodologies. Such design methodologies may be approached from two different starting points, both leading to optimum performances. On one side, since device efficiency is strongly dependent on the amount of power dissipated in the device itself \( P_{\text{diss}} \), a possible strategy consists in its minimization, obtained by a proper shaping of voltage and current waveforms aiming to avoid or minimizing overlapping regions. This waveform shaping can be realized by a proper output network design strategy, as in the Class-F approach [2–6], or by a switched active device operating condition, as the Class E [10–14]. On the other hand, a different approach may be attempted, trying to maximize the fundamental output voltage (or current) components, implying therefore higher output power \( P_{\text{out}} \) and efficiency. This aim can be obtained by loading the active device with a purely resistive fundamental load [22] and flattening the voltage (or current) waveform while approaching the device physical limitations, resulting in a potentially higher fundamental-frequency component (as will be clarified in the next section). The two above-mentioned strategies are substantially equivalent, as can be easily derived from power balance considerations, based on the following relation:

\[
P_{\text{in}} + P_{\text{DC}} = P_{\text{diss}} + P_{\text{out}}.
\]  

For a given power supplied to the active device (both from the DC bias \( P_{\text{DC}} \) and the RF input \( P_{\text{in}} \)), design methods devoted to the increase in device output power \( P_{\text{out}} \) or to the decrease in dissipated power \( P_{\text{diss}} \) in the active device itself seem to be equivalent. In fact both the above-mentioned approaches lead to the improvement of the device efficiency \( \eta = P_{\text{out}}/P_{\text{diss}} \) and require a proper “waveform engineering”, resulting in a careful selection of harmonic terminations.

To infer some useful design criteria for the output networks, it is necessary to make some simple consideration about the active devices used for microwave applications, usually field-effect devices. In fact, they can be effectively
treated as voltage-controlled current sources \cite{23, 24}, at least while operating in their active region. As a consequence, the resulting output current waveform is imposed by the controlling input voltage and, at least to the first approximation, does not depend on the output terminating impedances, that actually contribute to the output voltage wave shaping only. Thus, if the drain current waveform $I_d(t)$ is expressed in terms of its harmonic components $I_{d,nf_0}$:

$$I_d(t) = I_{d, DC} + \text{Real} \left( \sum_{n=1}^{n_{f_0}} I_{d, nf_0} \cdot e^{j2\pi nf_0 t} \right)$$ \hspace{1cm} (2)

the harmonic components $V_{ds,nf_0}$ of the drain voltage waveform can be simply expressed as:

$$V_{ds,nf_0} = Z_{nf_0} \cdot I_{d,nf_0},$$ \hspace{1cm} (3)

where $Z_{nf_0}$ is the actual impedance presented to the current source at the nth harmonic, resulting in a drain voltage waveform given by:

$$V_{ds}(t) = V_{ds, DC} - \text{Real} \left( \sum_{n=1}^{n_{f_0}} V_{ds, nf_0} \cdot e^{j2\pi nf_0 t} \right)$$ \hspace{1cm} (4)

thus suggesting the possibility of a voltage wave shaping both operating on the harmonic content of the current waveform $I_{d,nf_0}$ (choosing a proper input harmonic network) and/or on harmonic terminations $Z_{nf_0}$ (choosing a proper output network).

It is to note that the assumption above is valid to the first approximation only, and it is introduced for clarity; it can be however removed in actual designs, where a full nonlinear model for the active device and a nonlinear simulator are used, without major modifications in the result of the presented theory.

A second assumption however arises considering the number of frequency components that can be effectively controlled in a real design. On one hand, in fact, circuit complexity issues suggest the use of the minimum number of circuit idlers, given not only their chip area occupation but also availability and effectiveness of components’ models at high frequencies. On the other hand, the benefits that can be obtained by the control of a larger number of harmonic components do not justify such an increase. A reasonable and satisfactory compromise is in the control of the first two voltage harmonics (namely the second and third components), considering the higher ones effectively shorted by the prevailing capacitive behavior of the active device output. As a further justification of such an assumption, it should be noted that the control, up to the fifth harmonic component, has been implemented only at the low-frequency range \cite{9}, resulting more in higher circuit complexity than in a major efficiency improvement. The control scheme depicted in Figure 1 therefore represents a reasonable compromise among the various issues and constraints.

A further consideration regards the maximum output power condition for a given device. In Class A operation it can be obtained simultaneously by maximizing voltage and current swings \cite{22}, as schematically depicted in Figure 2. Such a condition can be easily extended to Class AB bias and it can be shown to be equivalent to a purely resistive loading of the controlled source, i.e. to the latter delivering active power only.

Because a resistive termination is the optimum load for output power maximization, the same holds for harmonic frequencies. In fact, complex

![Figure 1. Input and output terminating scheme of a multiharmonic manipulated PA.](image-url)
III. THE MULTIHARMONIC MANIPULATION THEORY: MATHEMATICAL STATEMENTS

On the basis of the assumptions in Section II, expression (4) can be explicitly rewritten utilizing second and third harmonic components only as:

\[
V_{ds}(t) = V_{ds,DC} - V_{ds,fo} \cdot \cos(2\pi f_o t) - V_{ds,2fo} \cdot \cos(2 \cdot 2\pi f_o t) - V_{ds,3fo} \cdot \cos(3 \cdot 2\pi f_o t)
\]  

(5)

Normalizing to the fundamental frequency component \( V_{ds,fo} \), eq. (5) becomes:

\[
V_{ds,\text{norm}}(\vartheta) = \frac{V_{ds}(\vartheta) - V_{ds,DC}}{V_{ds,fo}} = - \cos(\vartheta) - k_2 \cdot \cos(2 \cdot \vartheta) - k_3 \cdot \cos(3 \cdot \vartheta),
\]

(6)

where

\[
k_2 = \frac{V_{ds,2fo}}{V_{ds,fo}}, \quad k_3 = \frac{V_{ds,3fo}}{V_{ds,fo}}, \quad \vartheta = \omega_o t.
\]

(7)

The drain voltage waveform is constrained to swing within the range dictated by the device physical boundaries, i.e., the drain–source breakdown voltage \( V_k \) and the drain–source breakdown voltage \( V_{ds, br} \), where the gate–drain junction becomes forward-biased. It is therefore necessary that:

\[
V_k \leq V_{ds}(\vartheta) \leq V_{ds, br}.
\]

(8)

It can be observed that without the contribution of harmonic components, the maximum drain voltage amplitude in linear conditions is given by

\[
V_{ds,fo,\text{max}} = \min [V_{ds,\text{DC}} - V_k, V_{ds,br} - V_{ds,\text{DC}}].
\]

(9)

As previously mentioned, the target of the multiharmonic manipulation procedure is to obtain an increase in fundamental frequency voltage component over the case in which no harmonic components are present. This effect is obtained by means of a proper shaping of the voltage waveform, constrained to swing between the same physical limitations. Such a statement implies that the target is to obtain \( V_{ds,fo} \geq V_{ds,fo,\text{max}} \), which is equivalent, for the physical constraints, to the inequalities:

\[
V_{ds,\text{norm}}(\vartheta, k_2, k_3) \geq -1, \quad \text{if} \quad V_{ds,fo,\text{max}} = V_{ds,\text{DC}} - V_k,
\]

\[
V_{ds,\text{norm}}(\vartheta, k_2, k_3) \leq 1, \quad \text{if} \quad V_{ds,fo,\text{max}} = V_{ds,br} - V_{ds,\text{DC}}.
\]

(10a, 10b)

For the sake of simplicity, only the case represented by eq. (10a) will be discussed, since it is the most common situation, but an equivalent analysis can be performed for the case of eq. (10b).

In mathematical terms, the problem of eq. (10a) is equivalent to finding the values of \( k_2 \) and \( k_3 \) allowing for an increase in fundamental frequency voltage component over the not manipulated one while respecting the same physical limitations (eq. 10). Such an increase can be quantitatively evaluated by means of a Voltage Gain Function \( \delta(k_2, k_3) \), defined by:

\[
\delta(k_2, k_3) \equiv \frac{V_{ds,fo}}{V_{ds,fo,\text{max}}} = \frac{-1}{\min_{\vartheta}[V_{ds,\text{norm}}(\vartheta, k_2, k_3)]}.
\]

(11)
The resulting fundamental frequency voltage component can be therefore expressed as:

$$V_{ds,fo}|_{MHM} = \delta(k_2, k_3) \cdot V_{ds,fo,max}$$

(12)

The selection of optimum design points (i.e., values for $k_2$ and $k_3$ maximizing the fundamental frequency voltage component) therefore implies the study of the voltage gain function. In the recent past, the problem has been treated for the case of a harmonic manipulation based on the use of a single harmonic component, i.e., second ($k_3 = 0$, Class G) [20, 26, 27], or third ($k_2 = 0$ i.e., Class F) [28] harmonic components. In the present case, when both the second and third harmonic components are used ($k_2 \neq 0$, $k_3 \neq 0$, for extension Class FG), the mathematical treatment is quite long and unfortunately the results cannot be expressed in closed form. The surface of the voltage gain function $\delta(k_2, k_3)$ in the $k_2, k_3$ plane is given in Figure 3, while its contour plot is given in Figure 4. A clear maximum is visible for the voltage gain function, reaching the optimum zone for $k_2 < 0$ and $k_3 > 0$. In this case, fundamental component is in-phase with the third harmonic and out-of-phase with the second harmonic. It is to note that Class F operation corresponds to points lying on the negative side of the vertical axis ($k_3 < 0, k_2 = 0$), while Class G corresponds to points lying on the negative side of the horizontal axis ($k_2 < 0, k_3 = 0$). The more classical TL approach, imposing short-circuit terminations at harmonic frequencies, is represented by the origin ($k_2 = k_3 = 0$).

Basic considerations can be carried out regarding the sign of the $k_2$ and $k_3$ harmonic coefficients. If Class F or Class G operation is considered, a narrow range of $k_3$ and $k_2$ can be fruitfully utilized for harmonic manipulation, corresponding to the regions of the respective axes in which the voltage gain function is greater than unity. In both cases, such a condition corresponds to harmonic components out-of-phase (i.e., with opposite sign) with respect to the fundamental one [26–28], giving rise to a “flattening” of the resulting drain voltage waveform while it approaches the physical limitation of the device (as in Figure 5a for the Class F case). On the other hand, an in-phase combination results in a peaking effect on the voltage waveform, thus approaching the physical limitation for a lower fundamental frequency.
component and hence decreasing the maximum achievable fundamental frequency voltage amplitude, as shown in Figure 5b. If the waveform for the class G case (Figure 6) is considered, a further observation may be performed: while a flattening of the voltage waveform is effectively obtained when the drain current is at its maximum, on the other hand a peaking effect occurs in the remaining part of the cycle. On the device output characteristics, this effect leads to the operating point potentially entering the device breakdown region, with potential detrimental effects on device reliability. To account for the peaking effect obtained using even harmonic for the manipulation, a Voltage Overshoot Function $\beta(k_2, k_3)$ may be introduced, defined as:

$$
\beta(k_2, k_3) \equiv \max_{\phi} \frac{\mathcal{V}_{ds,\text{denorm}}(\phi)}{\max_{\phi} \mathcal{V}_{ds,\text{denorm}}(\phi)_{k_2=0, k_3=0}} \cdot \delta(k_2, k_3). \quad (13)
$$

As it can be easily inferred, $\beta(k_2, k_3)$ directly gives the amount of the overshoot for a given $(k_2, k_3)$ combination, and must be accounted for to avoid unwanted breakdown occurrence. The contour plot for the Voltage Overshoot Function is shown in Figure 7: the maximum values for such a function, $2.77 \leq \beta(k_2, k_3) \leq 3$, reside close to the region giving optimum values for the Voltage Gain Function, stressing its relevance in actual designs.

A further statement can be performed on the way in which the voltage waveform is flattened while approaching device physical limitations. If an ‘equiripple condition’ is imposed upon the voltage waveform, i.e., its minimum values are imposed to be equal, a simple equation linking the $k_2$ and $k_3$ values can be derived:

$$
k_3 = \frac{k_2^2}{4 \cdot (k_2 + 1)}. \quad (14)
$$

The use of expression (14) allows for an explicit representation for the Voltage Gain Function under the equiripple condition that is given by:

$$
\delta(k_2) = \frac{4 \cdot (1 + k_2)}{5 \cdot k_2^2 + 8 \cdot k_2 + 4}. \quad (15)
$$

The plot of such a function that is superimposed on the contour plot for the general Voltage Gain Function is shown in Figure 8. The maximum value for $\delta(k_2, k_3)$ in the equiripple condition is given by:

$$
\delta(k_2, \delta_{\text{max}}, k_3, \delta_{\text{max}}) = \frac{1 + \sqrt{5}}{2} \approx 1.62
$$

and it is obtained for the couple:

$$
[k_2, \delta_{\text{max}}, k_3, \delta_{\text{max}}] = \left[-1 + \frac{1}{\sqrt{5}}, \frac{3 \cdot \sqrt{5} - 5}{10}\right] \approx [-0.55, 0.17].
$$
Figure 7. Contour plot of Voltage Overshoot Function $\beta(k_2, k_3)$ in the $k_2$ and $k_3$ plane.

Such a maximum value is coincident with the absolute maximum for the Voltage Gain Function. On the other hand, a different approach may be attempted that tries to flatten as much as possible the voltage waveform (‘maximally flat’ condition), as suggested in [29], i.e., imposing null first and second derivatives on the waveform itself. Such a condition is a subset of the equiripple one and the resulting value for the voltage gain function is given by:

$$\delta(k_2, \text{maximally flat}, k_3, \text{maximally flat}) = \frac{3}{2} = 1.5$$

corresponding to

$$[k_2, \text{maximally flat}, k_3, \text{maximally flat}] = \left[ -\frac{2}{5}, \frac{1}{15} \right] = [-0.4, 0.067]$$

Figure 8. The Voltage Gain Function $\delta(k_2, k_3)$ under the equiripple condition.

hence leading to a suboptimum design.

In Figure 9 an example for voltage waveforms synthesized for three different conditions (maximum value of $\delta(k_2, k_3)$, maximum value of $\beta(k_2, k_3)$ and maximally flat) are reported.

IV. THE MULTIHARMONIC MANIPULATION THEORY: DESIGN STATEMENTS

The voltage harmonic shaping described in the previous section must now be related to the actual

Figure 9. Drain voltage waveforms under different conditions: maximum linear (dotted line): maximally flat (dot–dashed line); maximum of $\delta(k_2, k_3)$ (solid line): maximum of $\beta(k_2, k_3)$ (dashed line).
increase in power performances and to the output networks’ design. To this goal, let us briefly recall the rationale behind multi harmonic manipulation.

For a given device with its physical limits, a given maximum linear swing is allowed for the drain voltage (from eqn. (9)), whose time domain waveform is constrained to swing between the ohmic and breakdown regions. The intrinsic drain current is imposed by the drive level of the input waveform, thereby fixing its harmonic components. The maximum output power that can be obtained under such linear-operating conditions is simply given by the product of the maximum linear fundamental voltage component \( V_{ds,fo,max} \) times the drain current fundamental component \( I_{d,fo} \). Their ratio uniquely determines the load impedance at fundamental frequency \( Z_{fo} \) to be imposed, that is, on the basis of the discussion in Section II, a purely resistive termination

\[
R_{TL, opt} = \frac{V_{ds,fo,max}}{I_{d,fo}}, \quad (16)
\]

In this case harmonic terminations can be thought to be set to short-circuit ones, and the obtained design is the well-known TL strategy.

Starting from such a situation, and supposing that the harmonic components of the drain current are not influenced by their terminations (Section II), voltage harmonic components (second and third) can be added to the fundamental one according to their weights \( k_2 \) and \( k_3 \) computed in Section III. The result of such a wave shaping is a new voltage waveform with the same fundamental component, but with a reduced swing. The fundamental drain voltage component can be now increased by the factor \( \delta(k_2, k_3) \), to reach the device limitations. In this way, for the same drive level and with the same voltage swing, a higher fundamental frequency voltage component and therefore a higher output power is obtained.

Applying multiharmonic manipulation, the fundamental frequency voltage component is increased by the factor \( \delta(k_2, k_3) \) obtaining as indicated in eqn.12, here repeated for convenience:

\[
V_{ds,fo}|_{MHM} = \delta(k_2, k_3) \cdot V_{ds,fo,max} \]

and therefore the load to be imposed at fundamental frequency to obtain this goal is:

\[
R_{fo}|_{MHM} = \delta(k_2, k_3) \cdot R_{TL, opt}, \quad (17)
\]

Similarly, the harmonic terminations that have to be imposed at second- and third-order components can be computed by:

\[
R_{nfo}|_{MHM} = \delta(k_2, k_3) \cdot k_n \cdot \frac{I_{d,fo}}{I_{d,nfo}} \cdot R_{TL, opt}, \quad n = 2, 3. \quad (18)
\]

Fundamental frequency drain current component is, to the first approximation, unaffected by the resulting increase in the respective drain voltage component. Output performances are therefore increased by the same amount, i.e.:

\[
P_{\text{out, } MHM} = P_{\text{out, TL}} \cdot \delta(k_2, k_3) \quad (19a)
\]

\[
G_{\text{out, } MHM} = G_{\text{out, TL}} \cdot \delta(k_2, k_3), \quad (19b)
\]

\[
\eta_{d,MHM} = \eta_{d,TL} \cdot \delta(k_2, k_3). \quad (19c)
\]

Expression (17) gives the optimum fundamental frequency termination and in its simplicity reveals a potential source of error while performing PA design. A widely used procedure to investigate the power performances of a given device is to measure its load–pull contours. Load–pull systems are nowadays becoming extremely sophisticated, giving the possibility to perform load–source pull measurements not only at fundamental but also at harmonics. The usual procedure, in the case of harmonic load pull, consists in finding the optimum fundamental frequency termination for fixed values of harmonic loads. Once such a value is determined, it is held fixed and the harmonic loads are varied until an optimum value for them is found. On the basis of the theory in the previous section, such a combination of loads is not the optimum one, since the fundamental frequency load without (or for a fixed) harmonic tuning is not the same that can be obtained varying harmonic loads. A correct load–pull procedure should vary harmonic load together with the fundamental one to find the global optimum combination. On the other hand, eq. (17) may be used to find a step-by-step procedure, starting from the TL case.

V. HARMONIC GENERATION MECHANISMS AND DRAIN CURRENT WAVEFORMS

In this section the problem of the proper current harmonic generation will be addressed. In
fact, since passive terminations only can be obviously employed, the properly phased voltage harmonic components must result from eq. (3), i.e., starting from the output drain current harmonic components and choosing suitable terminations. Different approaches can be explored to obtain the proper phase relationships among the drain current harmonic components, and will be briefly examined in the following.

A first possibility consists of the use of the output clipping phenomena, i.e., in the generation of current harmonic components by means of hard device nonlinearities as the pinch-off and the input gate–source junction forward conduction. Since this phenomenon is related to the input drive level ($P_{\text{in}}$) and to the selected bias point, it implies a proper selection of the active device operating conditions. If a simple sinusoidal drive is used as input signal, the resulting drain current is simply a truncated sinusoid, whose conduction angle ($\vartheta_c$), defined in Figure 10a, completely determines the resulting phase relationships among current harmonics and reveals some important properties. By a simple Fourier transformation, the harmonic components can be computed as plotted in Figure 10b. As it is possible to note, for conduction angles ranging from Classes A to B, the second harmonic component $I_2$ is always in-phase and the third one $I_3$ is always out-of-phase with respect to the fundamental component $I_1$ (i.e., having the same and opposite sign, respectively). As a consequence, the direct application of the multiharmonic manipulation procedure described beforehand, i.e., with purely resistive harmonic loads, is not allowed. Only a Class F design is therefore directly applicable [28], although it becomes deleterious for Class C bias conditions. Even if a second nonlinear phenomenon (i.e., the input diode forward conduction ($\vartheta_b$)) is encountered, the picture modifies only slightly, as shown in Figure 11.

As a first remark, because simple resistive harmonic terminations are not useful, complex ones are required, also at fundamental, partially vanishing the improvement due to the multiharmonic manipulation by the fundamental reactive power produced. In this case, a simple suitable design criterion is obtained choosing the harmonic terminations as dictated by the high-frequency Class E approach, while paying at least a higher overshoot factor $\beta \approx 3.65$ [30].

A second opportunity consists of using the effect of device input nonlinearities. A Volterra analysis of the input circuit [26, 31] shows that the main contribution to the harmonic generating mechanism at the device input is given by the nonlinear input capacitor $C_{gs}$, thus confirming the numerical and experimental results in [15, 16]. If reasonable drive levels are considered, the control voltage $V'_{gs}$ exhibits a major second harmonic content, leading to an asymmetrical gate–source voltage waveform, as depicted in Figure 12. To avoid this effect, which is often considered as detrimental, the input harmonic terminations are frequently set to short-circuit values [15, 17] or compensated by means of a counteracting nonlinearity [16]. Nevertheless, major improvements of power performances are obtained if such a second harmonic input voltage component is used to implement the technique described beforehand. In fact, the input signal nonlinear distortion implies the generation of a second harmonic gate voltage.
component that is ‘out-of-phase’ with respect to the fundamental one and therefore usable for the generation of output current components with the same phase relationship. Moreover, also a third harmonic component ‘in-phase’ with the fundamental one is generated, suitable for a Class FG multiharmonic manipulation.

Up to now therefore, while nonlinear output clipping phenomena determine a ‘wrong’ phase relationship among current harmonics, input nonlinearities effectively act in a reverse direction, generating both the second and third harmonic components with the proper phasing. These two counteracting effects cooperate in a very complex way in real devices. On the other hand, it is clear that the input nonlinearities dominate at moderate drive levels, while output clipping phenomena should prevail for higher levels. Such a behavior strongly depends on biasing conditions, since the latter fix the drive level at which physical limitations are incurred: roughly speaking, the closer the bias point to the class-A reference, the higher will be the drive level at which the counteracting output harmonic generation prevails.

Another aspect that must be considered is related to the amplitudes of the ratios among the voltage harmonic components, i.e. the values of $k_2$ and $k_3$. In fact, even if the phase relationships are correct, the values of $k_2$ and $k_3$ are related to the drain current harmonic components and to the harmonic load resistances by eq. (18). While the amplitude of the harmonic components increase with the input drive signal, the harmonic load resistances are upper limited by the output device resistance value $R_{ds}$. Such a behavior is demonstrated for a typical power stage in Figure 13, where the relative amplitudes of both the second and third harmonic drain voltage components with respect to the fundamental one ($k_2, k_3$) are plotted as a function of the input power for a fixed bias point and loading (both input and output).

A third different approach can be however attempted, even if it is more unusual. Since the drain voltage waveform is built from the drain current harmonic components, resulting from eq. (3), to obtain the proper phase and amplitude relationships, proper drain current harmonic components could be generated. To the latter goal, the input nonlinearities can be fruitfully employed, as noted earlier, but this is not the unique possibility. In fact, a preshaped waveform may be fed to the input of the power stage, containing already the correct phasing between its harmonics.

Even if it is possible to analyze the best input drive waveform for each harmonic strategy, as in [21] for Class F design, that could be unfeasible due to its hard implementation and too sensitive to the active device input model assumed,
Figure 13. Class FG design: normalized amplitude of drain voltage 2nd (dashed line) and 3rd (solid line) harmonics.

more practical consideration leads to the analysis of realistic and easy to implement cases only. For instance, a suitable waveform is obviously a square waveform, that may be easily obtained using a class-F amplifier [28] as the driver stage. In this case, the analysis may start directly from the drain current waveforms, that in the following will be assumed as a square waveform, as a truncated sinusoid (as the reference case before examined), and finally as a quadratic waveform, to take into account a more realistic active device pinch-off nonlinear behavior, as suggested by [32]. Through a Fourier analysis on the three current waveforms, the regions in which only output purely resistive harmonic manipulation is allowed can be evidenced, as reported in Table I. The drain current conduction angle \( \vartheta_c \) is to be considered, for the three cases, as the region where the drain current is nonzero, and it is coincident with the duty cycle for the square waveform.

Using a piecewise linear simplified model for the active device, for the regions of Table I, the expected improvements in terms of output power and drain efficiency can be evaluated through eq. (19). The theoretical output power (normalized to a standard Class A design) and the drain

### TABLE I. Drain Current Circulation Angles Allowing Class FG Approach

<table>
<thead>
<tr>
<th>Current Waveform Model</th>
<th>Class FG</th>
</tr>
</thead>
<tbody>
<tr>
<td>Truncated sinusoid</td>
<td>Never possible</td>
</tr>
<tr>
<td>Quadratic</td>
<td>( 6.06 &lt; \vartheta_c &lt; 2\pi )</td>
</tr>
<tr>
<td>Square</td>
<td>( 4.18 &lt; \vartheta_c &lt; 2\pi )</td>
</tr>
</tbody>
</table>

Figure 14. Normalized output for a Tuned Load PA (dotted) and Class FG PA(solid): (a) quadratic and (b) square drain current waveform.

efficiency for a TL and Class FG amplifiers, are depicted in Figures 14 and 15, respectively.

It is to be noted that the theoretical purely resistive multiharmonic manipulation seems to be useful only for a narrow range of the drain current circulation angle \( \vartheta_c \); moreover, the efficiency improvements could not be satisfactory, i.e. for a high efficiency design could be more appropriate choice values of \( \vartheta_c \) closest to Class B bias condition, obtaining higher efficiency values. As a consequence, the optimum design is a tradeoff all the above-mentioned phenomena.

In summary, many different solutions seem to be available, using output and/or input manipulations to obtain significant improvements over the classical TL performances. Obviously, a combined action both at input and output ports could represent the best solution, depending on the acceptable growth in circuit complexity. Moreover, especially for the simplified analysis based on the three different driving signals above, the reported results represent only a first-order approximation, while the effects of the input and output nonlinearities are not accounted for. In any case, a more accurate analysis based on a full nonlinear model of an actual device, including all the sources of nonlinear behavior, demon-
VI. SAMPLE REALIZATION AND MEASURED PERFORMANCES

To demonstrate the effectiveness of the proposed multiharmonic manipulation strategy for high efficiency design, two power amplifiers have been designed and realized to operate at $f_0 = 5$ GHz (fundamental frequency), namely a Tuned Load reference stage, and a Class FG amplifier that utilizes both the second and third harmonic voltage tunings.

The device used is a medium power MESFET by Alenia Marconi Systems with a Class AB bias condition ($I_q = I_{dss}/3 \approx 80\text{mA}; V_q = 5\text{V}$) and 1 mm gate periphery. The device has been modeled in-house by a full nonlinear model, whose topology is depicted in Figure 16, extracted using multibias S-parameter and pulsed-dc measurements [33]. For the two designs, the choice of the fundamental-frequency output termination has been optimized by means of the technique in [34] and an input matching network has been synthe-

![Figure 15](image1)

Figure 15. Drain efficiency according to a simplified model for a Tuned Load PA (dotted) and Class FG PA(solid): (a) quadratic and (b) square drain current waveforms.

![Figure 16](image2)

Figure 16. Device nonlinear model.
sized to get maximum input power transfer at large-signal (conjugate large-signal input match) and to generate, if necessary, the drain current harmonic components with the appropriate phase relationships, thus applying the considerations performed in the previous section, as will be briefly explained in the following.

In particular, for the TL design the input capacitor $C_{gs}$, whose terminal voltage directly controls the output current source, is nearly short-circuited for the harmonic components at $2f_0$ and $3f_0$, while for the Class FG amplifier, it is properly terminated to allow for the multi-harmonic manipulation, i.e. to increase the input harmonic content, as depicted in Figure 17. The two solutions have been obtained loading the input circuit by an almost open (TL approach) or short-circuit (Class FG approach) external termination at $2f_0$ and $3f_0$. Such loads are transformed by the parasitic network in the proper short (TL approach) or open-circuit (Class FG approach) loading across $C_{gs}$.

In the same way, the output matching networks are designed following different criteria for each amplifier. For the TL amplifier, the output network actually shorts the harmonic components of the intrinsic drain current at $2f_0$ and $3f_0$, hence obtaining an almost-sinusoidal voltage output waveform. Contrarily, for the Class FG PA, the second and third drain current harmonics are terminated to shape the drain voltage waveform as described in Section V. With reference to Figure 16, and accounting for the consideration in the previous section, the values to be synthesized at the intrinsic drain terminals at $f_0$, $2f_0$ and $3f_0$ are summarized in Table II for the TL and FG PA. It is to be noted for the latter that the proper relative phase of the second harmonic voltage component with respect to the fundamental one is obtained by means of both the input and output second harmonic terminations. This is due to the fact that, in this case, a purely resistive output harmonic manipulation is not sufficient and in the same manner, input nonlinearities only cannot assure the proper drain current phase relationships. Both effects have been therefore utilized. The Class FG design has been obtained by synthesizing a purely resistive fundamental load and complex harmonics thus assuring the proper drain voltage components phase relationships, as shown in Figure 18. The values of the optimum terminations at the extrinsic device terminals are summarized in Table III.

To synthesize the external loads in Table III, the distributed approach schematically drawn in Figures 19 (TL design) and 20 (Class FG design) has been followed, for the input and output networks, respectively. Moreover, the TL output network is realized by means of two stubs ($\lambda/8$ open-circuit stub and $\lambda/6$ short-circuit stub) controlling both the second and third harmonic terminations, respectively. The fundamental load is synthesized by a standard LC cell. The Class FG output network is simpler because it is obtained when started by a $\lambda/12$ short-circuit stub that

**Figure 17.** Simulated intrinsic gate voltage components at 1dB compression point for TL and Class FG-designed amplifiers.

**Table II. Intrinsic Drain Termination for the Realized TL and Class FG Amplifiers**

<table>
<thead>
<tr>
<th>Freq (GHz)</th>
<th>Tuned Load</th>
<th>Class FG</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>26.5 + j0.1</td>
<td>49.7 + j3.7</td>
</tr>
<tr>
<td>10</td>
<td>2.4 + j0.1</td>
<td>22.7 + j131.1</td>
</tr>
<tr>
<td>15</td>
<td>2.2 − j0.1</td>
<td>8.7 + j11.8</td>
</tr>
</tbody>
</table>

PA.
Figure 18. Simulated drain voltage and current harmonic components at 1dB compression point for TL and Class FG-designed amplifiers.

### TABLE III. External Input and Output Impedances for the Realized TL and Class FG Amplifiers

<table>
<thead>
<tr>
<th>Freq(GHz)</th>
<th>Tuned Load</th>
<th>Class FG</th>
<th>Tuned Load</th>
<th>Class FG</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Input</td>
<td>Output</td>
<td>Input</td>
<td>Output</td>
</tr>
<tr>
<td>5</td>
<td>14.4 + 25.9j</td>
<td>22.3 + 6.1j</td>
<td>14.5 + 25.8j</td>
<td>32.6 + 20.5j</td>
</tr>
<tr>
<td>10</td>
<td>42.9 – 223.4j</td>
<td>1.6 – 4.6j</td>
<td>0.5 + 14.8j</td>
<td>3.4 + 68.1j</td>
</tr>
<tr>
<td>15</td>
<td>23.9 – 130.4j</td>
<td>1.3 – 7.2j</td>
<td>2.0 + 26.8j</td>
<td>42.9 – 469.2j</td>
</tr>
</tbody>
</table>

Figure 19. Tuned Load design network criteria.
controls third harmonic component and an LC cell to control fundamental and second harmonic impedances. Biasing voltages have been applied through the rf signal connectors.

Simulated results are reported in Figures 21 and 22, where the drain voltage waveforms and the corresponding I–V load curves, computed at -1dB gain compression point, are indicated as obtained by a full nonlinear simulator (HP-MDS). It is to be noted that the use of a second harmonic component for the Class FG PA produces, as expected, a peaking effect on the drain voltage waveform, whose value can be predicted by eq. (13) and must be accounted for in order to avoid device breakdown.

In Figure 23 the layouts of the two PAs are reported, as realized in hybrid form on Alumina substrates. Plots of measured output power and PAE as functions of the input drive at 5 GHz are shown in Figures 24 and 25, respectively. As expected, the use of the multiharmonic manipulation significantly improves the PAs performances. In particular, for an input drive level of 18.3dBm, a maximum PAE was obtained for a Class FG PA, with measured output power and PAE levels of 25.6 dBm and 60%, respectively, corresponding to a drain efficiency of 73.7% (Table IV). The remarkable improvements in output power and PAE are synthesized in Table II, where a measured improvement factor of 1.43 for Class FG with respect to TL amplifier is reported. This figure is not far from the theoretically expected value (1.56).

A final statement has to be discussed to clarify an ambiguous problem. Harmonic tuning strategies are often referred as detrimental approaches if the linearity of the stage has to be addressed.
To investigate this aspect, intermodulation two-tone tests have been performed on the two amplifiers, injecting two equal-amplitude input signals 50 MHz apart. The measured results, plotted as a function of the OBO are shown in Figure 26. As can be noted from this figure, for the Class FG approach the generation of an input second harmonic [35] actually decreases the IM distortion, thus suggesting that a multiharmonic manipulation strategy may allow for not only an increase in power performances, but also a tradeoff with linearity if IMD levels are of concern. The final consideration is with regard to the presence of a sweet spot, i.e., a null IMD value that is basically unchanged by the multiharmonic manipulation strategy. It seems to depend on the selected bias point, but further investigations are necessary.

VII. CONCLUSIONS

The potentials of multiharmonic manipulation PAs design, using a proper generation and termination of the second and third harmonics, have been investigated. Moving from the TL approach to a harmonic manipulated (Class FG) solution, major improvements were demonstrated in terms of output power, gain and efficiency, by a simplified analysis that can be performed for the multiharmonic manipulation procedure.

Moreover, two 5 GHz PAs have been designed, realized in hybrid form and measured and their performances confirm the remarkable improvements that can be obtained by means of the proposed multiharmonic manipulation strategy, while the IMD performances are improved or at least unaffected by this approach. In particular, the Class FG design approach demonstrated a
Figure 25. Measured PAE.

<table>
<thead>
<tr>
<th>Port [dBm]</th>
<th>PAE [%] Measured</th>
<th>PAE [%] Expected</th>
</tr>
</thead>
<tbody>
<tr>
<td>TL</td>
<td>25.0</td>
<td>42</td>
</tr>
<tr>
<td>FG</td>
<td>25.6</td>
<td>60</td>
</tr>
</tbody>
</table>

Table IV. Performances of TL vs Class FG PA

Figure 26. IMD measurements vs Output Back-Off (OBO).

Theoretical improvement factor of 1.63 in terms of efficiency, output power and gain, over the standard TL approach.

ACKNOWLEDGMENTS

The Advanced Research Technology Group of Alenia Marconi Systems is deeply acknowledged for their support in realizing and measuring the hybrid amplifiers.

REFERENCES

Multiharmonic Manipulation for Power Amplifiers


BIOGRAPHIES

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