RF versus Microwave High Efficiency PA Design

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Abstract

In this paper a comparison between classical Class E and new Class FG power amplifier design strategies are presented, to evaluate which method is best suited for high efficiency performance when operating frequency lies in the low microwave range. Fundamental design guidelines are briefly discussed and the Class E strategy is critically revisited with a frequency domain approach, stressing the differences with the Class FG. Finally, in order to validate the analysis, measured performances of a realised Class FG PA will be presented and compared with the simulated results of a Class E PA. For the same operating conditions, Class FG amplifier exhibits superior performances, with a maximum PAE of 59.8% (measured 58.8%) and a saturated output power of 25dBm, compared to 48% and 24dBm of Class E. Moreover, the 1dB compression point is 3dB higher for the Class FG.

Introduction

Over the last years, many design methodologies have been proposed in literature for microwave power amplifiers (PA's), with the common target to increase the efficiency performances. The theoretical approaches developed range from the simple Tuned Load (TL) strategy [1], useful to get basic design information, to more complex harmonic-tuning, as the Class F amplifier suggested by Snider [2].

In the first case (TL), the output voltage harmonic components, generated by the active device nonlinearities, are simply terminated on short-circuit loads. In the second case, the idea is to "control" these harmonics by proper terminations to obtain output voltage and current waveforms minimising the dissipated power in the active device, and therefore increasing the conversion efficiency. In the classical Class F approach, the output voltage waveform is squared by choosing an ideal output network shorting the even harmonic components and opening the odd ones [2]. The design is completed synthesising an input network fulfilling the conjugate (large signal) matching conditions at fundamental frequency. The above mentioned design strategies have been successfully implemented up to a few GHz but exhibit a different behaviour increasing the operating frequency, due both to harder control of the higher harmonic terminations and to the active device output capacitive behaviour, practically shorting higher components and not allowing therefore the desired waveshaping. Class F approach becomes a sub-optimum design, as demonstrated in [3, 4]. Recently the authors proposed a new design approach for practical Class F amplifiers [3], evidencing some failures of the Snider theory and giving some useful design guidelines. Moreover, the harmonic generating mechanism has been clarified together with the role of the phase relationships among the various harmonic components. Generalising the approaches in [3, 4], a new harmonic tuning strategy (Class FG) has been proposed in [5].

On the other hand, if the PA design is to be performed at RF or at the low end of the microwave frequency range, a common solution is in the adoption of completely different approaches, the most famous of which is the Class E strategy, originally developed by N.O. and A.D.Sokal in 1975 [6]. The method is based on an active device acting as a switch, so that either the drain voltage or the drain current is ideally zero, therefore nulling the dissipated power and resulting in high efficiency performance. Using a time domain approach, the design equations, i.e. the expressions for the output network element values, can be determined in closed form. Nevertheless, it is to note that the stage realised with this approach is properly a DC-RF converter rather than a power amplifier, since there is not a clear relationship between the input (Pin) and output (Pout) signals: no information on the stage large signal gain is available beforehand and the gain itself cannot be a specification but only the result of the design. To obtain the proper driving signal for the Class E, the solution proposed in the past was the realisation of a cascade of Class F (driver) and Class E (final) PA's, defining the transfer function of the complete amplifier. Recently, single stage power amplifiers, driven with a sinusoidal input signal and designed using the Class E approach, have been proposed [7]. In both solutions, the input network of the stage is synthesised to impose conjugate matching and nice performances both in terms of output power and efficiency have been obtained for the stages operating in the low microwave frequency range.

From the discussion above, it is not clear therefore which method is best suited for high efficiency operation, above all if the operating frequency lies in the low microwave range. In this contribution the basics of the harmonic manipulation approach (HMA) will be briefly recalled and the Class E strategy will be revisited under the proposed theory, utilising a frequency domain approach and stressing the differences with the Class FG. Finally, in order to validate the analysis, measured performances of a realised Class FG PA will be presented and compared with the simulated results of a Class E PA.

Design approaches: Class FG *vs* Class E

The starting hypothesis of the HMA is to assume that the output voltage waveform is shaped through a proper choice of harmonic output terminations, i.e.

$$V_{ds}(t) = V_{ds,DC} - \Re e\left(\sum_{n=1} V_{ds,nf_0} \cdot e^{j2\pi n f_0 t}\right) \qquad V_{ds,nfo} = Z_{nfo} \cdot I_{ds,nfo} \tag{1}$$

where $V_{ds,DC}$ is the dc component and $V_{ds,nfo}$ are the harmonic components at frequency nf_o and $Z_{L,nfo}$ is the termination imposed to the *n*-th harmonic current component at the intrinsic drain terminals. Considering the ac component, normalised to the fundamental amplitude, a normalised version of the drain voltage can be used, defined as

$$V_{ds,norm}(\vartheta)=\frac{V_{ds}(\vartheta)-V_{ds,DC}}{V_{ds,fo}}=-\cos(\vartheta)-k_2\cdot\cos(2\cdot\vartheta)-k_3\cdot\cos(3\cdot\vartheta)\qquad \vartheta=\omega_o t \tag{2}$$

where

$$k_2=\frac{V_{ds,2fo}}{V_{ds,fo}}\qquad\qquad k_3=\frac{V_{ds,3fo}}{V_{ds,fo}} \tag{3}$$

For the TL case, the maximum achievable fundamental amplitude is $V_{ds,fo,max}$=min[$V_{ds,DC}$-V_k, $V_{ds,br}$-$V_{ds,DC}$] and the voltage waveform swings between the device physical limits, $V_k \le V_{ds}(\vartheta) \le V_{ds,br}$ (where V_k is the drain knee voltage and $V_{ds,br}$, is the drain-source breakdown voltage). HMA solves the problem of finding proper values of k_2 and k_3 allowing an increase in fundamental frequency voltage component over the TL case and maintaining the waveform swing within the device physical limitations.
The results of the mathematical treatment of the problem, can be summarised defining a ***Voltage Gain Function*** *$\delta(k_2,k_3)$* as the increase in fundamental frequency voltage component over the maximum linear one. The resulting fundamental frequency voltage component can be expressed as

$$V_1\big|_{HM}=\delta(k_2,k_3)\cdot V_1\big|_{TL} \tag{4}$$

As a consequence, the improvements of the PA's performances with respect to the reference TL amplifier, can be easily expressed as:

$$P_{out,HM}=P_{out,TL}\cdot\delta(k_2,k_3)\qquad\qquad \eta_{d,HM}=\eta_{d,TL}\cdot\delta(k_2,k_3) \tag{5}$$

In fig. 1 the contour plot of *$\delta(k_2,k_3)$* is presented in the k_2-k_3 plane, where the importance of the voltage components phase relationships with respect the fundamental one is evidenced by the sign of k_2 (<0, i.e. out-of-phase) and k_3 (>0, i.e in-phase). When this situation occurs, i.e. when the output current components have a proper phase relationship, a purely resistive load can be chosen, whose values are:

$$R_{load,1,FG}=\delta_{FG}(k_2,k_3)\cdot R_{load,1,TL}$$

$$R_{load,2,FG}=\delta(k_2,k_3)\cdot k_2\cdot\frac{I_{ds,1,TL}}{I_{ds,2,TL}}\cdot R_{load,1,TL}\qquad R_{load,3,FG}=\delta(k_2,k_3)\cdot k_3\cdot\frac{I_{ds,1,TL}}{I_{ds,3,TL}}\cdot R_{load,1,TL} \tag{6}$$

so maximising the active power delivered from the device to the fundamental load.
The correct phase relationships between output current harmonics can be obtained through a proper choice of the input harmonic loads, exploiting the input nonlinear behaviour of the active device [8].
As previously outlined, the Class E approach is based on the idea that the active device acts as an ideal on/off switch, and the design criteria is to shape the output current and voltage waveforms to prevent their overlap, so minimising the power dissipation on the device also during the switch transitions and increasing therefore the efficiency performance. Typical Class E waveforms meeting the high-efficiency requirements are shown in fig. 2, while the scheme of the Class E PA is reported in fig. 3: the input network fulfils the conjugate match condition and the output network is designed according to [9], inferred through a time domain approach.
If a frequency domain analysis is carried out [10], it is possible to derive the phase and amplitude relationships between voltage harmonic components, i.e. the k_2 and k_3 quantities (see eqn 3), which can be compared with the same quantities defined for Class FG strategy. The results of this approach are summarised in Tab.1 and graphically shown in Fig. 4. Moreover, the fundamental output load inferred both with a time domain and a frequency domain approaches is given by

$$Z_1=\frac{0.28015}{\omega C_s}e^{j49.05^\circ} \tag{7}$$

where ω is the operating frequency and C_s the output capacitance of the transistor. From the results obtained, it can be noted that the voltage harmonic to fundamental components ratio have basically the same relationships both for Class FG and Class E approaches, even if for Class E the voltage phase differences are not exactly 0 or 180 degrees (as assumed for ideal Class FG approach).
The difference between the two approaches are in the fundamental load phase values, that is 0 (pure resistive load) for Class FG while is 49.05° for Class E. This means that the active power delivered from the device to the load is not maximum (decreased by a factor given by $\cos\phi \approx 0.66$). Nevertheless, this detrimental effect is partially compensated by a fundamental voltage amplitude higher for Class E than for Class FG.
On the other hand, the higher improvements obtainable with Class FG approach imply a more complex circuitry since an input harmonic manipulation is mandatory to obtain the proper output current phase relationships. It is possible to demonstrate in fact that utilising the input active device nonlinear behaviour, the input harmonic loads can be properly designed to manage the phase relationships, allowing purely resistive output loads [8].

Experimental Results

Two PA's have been designed to validate and explore the conclusions above, the first one utilising a Class FG approach and the second following a Class E strategy. Both amplifiers are single stages using a medium power GaAs MESFET (0,5 μm gate length) with 1mm gate periphery and interdigitated structure (10x100 μm) by Alenia Marconi Systems. A class-AB bias, corresponding to 30% of $\mathbf{I_{dss}}$ has been selected with 5V of drain voltage. Both the PA's have been designed at a fundamental frequency of 5 GHz. The scheme of Class E and the layout of the realised Class FG PA's are shown in fig. 3 and 5. In both the amplifiers, the input networks have been synthesised to realise the fundamental frequency matching. Higher harmonics are loaded by short-circuit terminations for the class E according to the criteria of minimising the active device conduction time.

Regarding the output networks, for the Class FG PA the optimum purely resistive fundamental load has been synthesised while the other harmonic output loads have been controlled according to the proposed methodology, i.e. trying to realise the proper values of k_2 and k_3 (eqn 9). On the other hand, the Class E output network has been synthesised according to Sokal equations [9].

The resulting voltage harmonic components amplitude and phase relationships are shown in fig. 6, where the input power is swept from linear to saturation regimes. It is possible to note that the phase and amplitude relationships of the voltage components are quite similar; moreover, while the output current waveforms are comparable, the maximum value of the Class E output voltage is higher than in the case of Class FG, stressing a higher fundamental component, as predicted in the previous sections.

Finally, the simulated output power and drain efficiency for both the amplifiers are shown in fig. 7 together with the measured performances for the realised Class FG PA. It can be noted that the Class FG amplifier assures both a higher output power and a higher drain efficiency, as forecasted in the previous discussion.

Conclusions

In this paper the classical Class E and the new Class FG power amplifier design strategies have been discussed, inferring useful design guidelines and stressing their differences, especially when the operating frequency lies in the low microwave range. A realised 5GHz Class FG amplifier has been compared with a simulated Class E in the same operating conditions, showing a maximum measured PAE of 58.8% (simulated 59.8%) and a saturated output power of 25dBm, compared with the simulated values respectively of 48.1% and 24dBm of the Class E.

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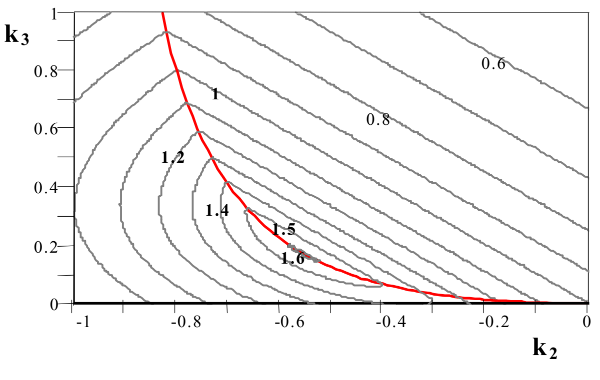

Fig. 1: The voltage gain function contour plot.

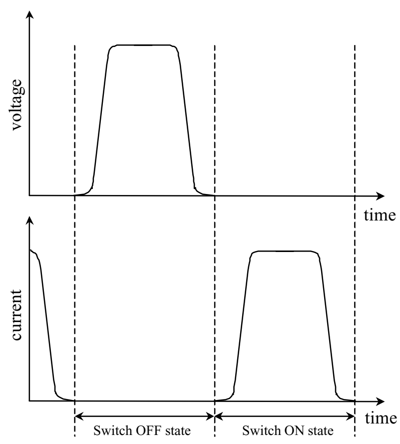

Fig. 2: Ideal Class E output waveforms.

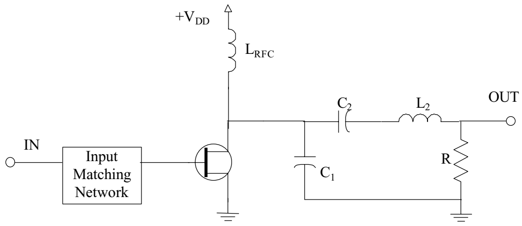

Fig. 3: Scheme for the Class E power amplifier.

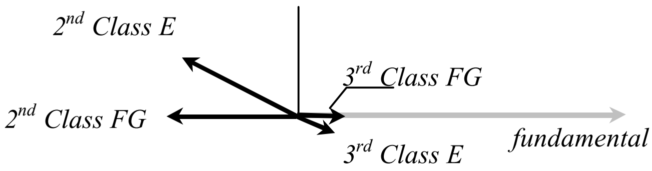

Fig. 4: Voltage harmonic components for Class FG and Class E PA's

Tab. 1

	$\lvert V_2\rvert/\lvert V_1\rvert$	$\angle(V_2/V_1)$	$\lvert V_3\rvert/\lvert V_1\rvert$	$\angle(V_3/V_1)$
Class FG	0.553	3.142	0.171	0
Class E	0.517	2.662	0.136	-0.431

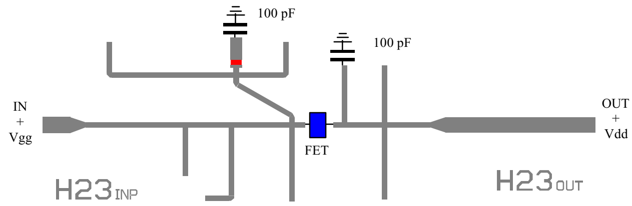

Fig. 5: Layout of designed Class FG amplifier.

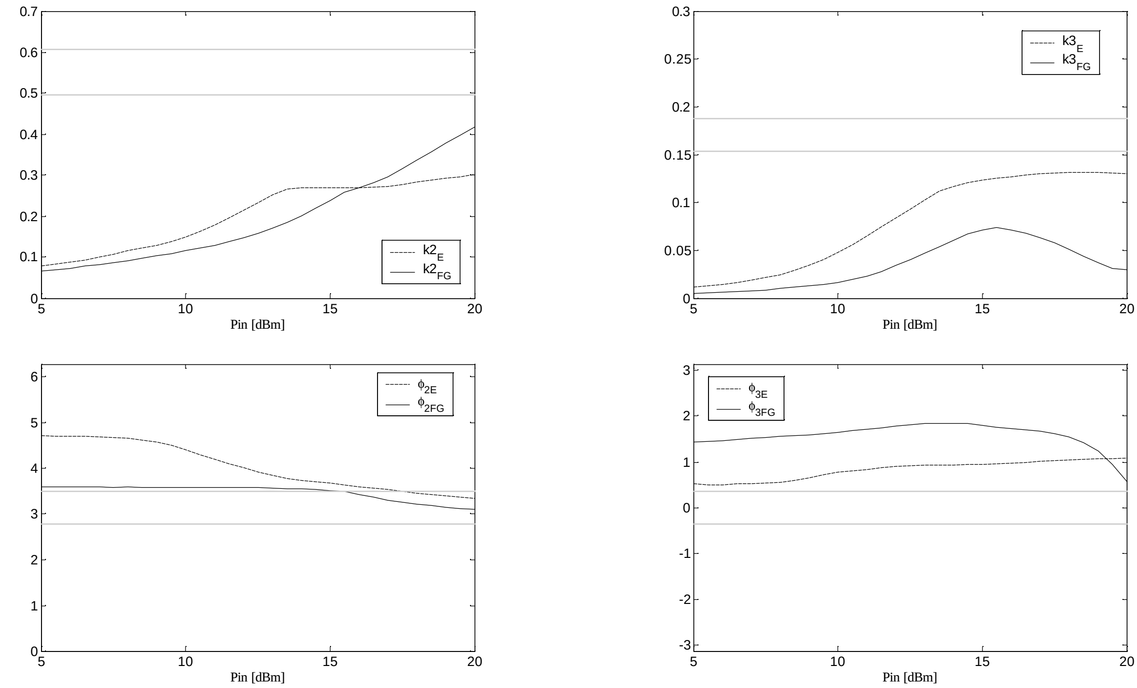

Fig. 6: Amplitude and phase relationships versus input power P_{in}, for Class E and FG PA's.

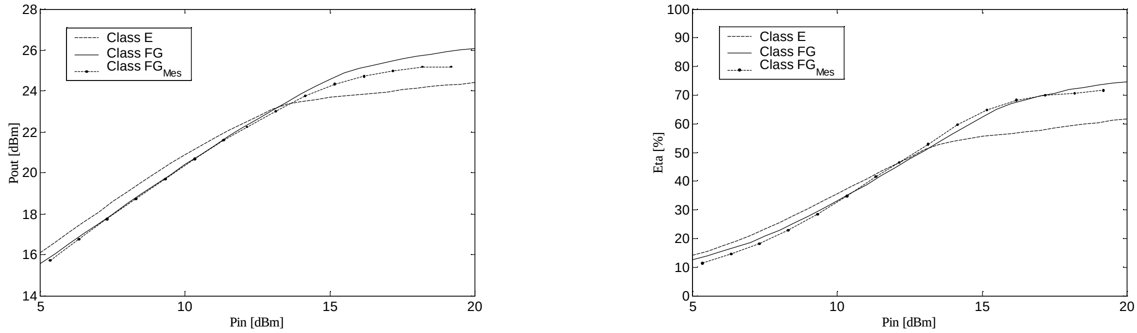

Fig. 7: Power amplifiers performances: (a) output power; (b) drain efficiency.