Simple Expression for Estimating the Switch Peak Voltage on the Class-E Amplifier With Finite DC-Feed Inductance

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Abstract—The class-E power amplifier is widely used due to its high efficiency, resulting from switching at zero voltage and zero slope of the switch voltage (i.e. ZVS and DZVS). This paper presents an analytical expression of the gain between the DC input voltage and the peak switch voltage on an ideal class-E power amplifier (PA) with finite dc-feed inductance, ZVS and DZVS operation. This expression is verified by simulations, and it is evaluated by experimental results at a switching frequency of 10.24MHz. Considering the results (simulated and experimental), the maximum error was 10.2%.

Keywords—Class-E, power amplifier, power efficiency, wireless power transfer.

I. INTRODUCTION

The Class-E power amplifier (PA) has been extensively studied and many different aspects of it has been analyzed. In particular, the ideal Class-E PA (i.e. Fig. 1(a)) with a finite dc-feed inductance instead of an RF-choke in a Class-E PA has been explored [1]–[5]. For the same supply voltage, output power and load, using finite dc-feed inductance has significant benefits [6]: more efficient output matching network, implementation in low-voltage technologies and higher frequency of operation. The published papers about this PA can be categorized, considering the design approach, in two main groups: based on analytical equations [3], [5], [6] and based on iterative procedures [1], [4]. Further, when the switch on-resistance and the inductor resistance are taken into account, the class-E PA solution results in complicated nonlinear analytic equations that must be solved numerically [7], [8] or in iterative design procedures even more lengthy and complex [9], [10]. Furthermore, the optimum operation of the class-E PA occurs at the non-nominal operation (i.e. without ZVS and DZVS) [10]. An intermediary solution was proposed in [11], they used the analytical solution of the ideal class-E PA equations as the first point of an iterative procedure for solving the optimization of the PA.

Considering the Class-E PA presented in the Fig. 1(a), the peak value of the switch voltage should be lower than the breakdown voltage of the transistor. The solutions for alleviating this problem include [7]: parallel structures and power combining, cascade switching, combining with class-F/F⁻¹, and designing for sub-optimum or off-nominal operation of the class-E PA. The published design methodologies for nominal or optimum operation that consider the switch breakdown voltage (V_{Sm}) involves hard simulation work [10] or numerical method solution of non-linear equations [7]. On the other hand, in [11] was included the V_{Sm} in an analytical design set, this set was divided in specification gains and circuit element gains as is illustrated in the Fig.2(a). The specification gains were used to calculate the design space of the class-E PA based in the inputs and outputs constrains, and the circuit element gains were used for calculating the circuit components. All of these gains are analytic functions of the input variables, therefore the design set can be implemented and calculated in any math software for analyzing all the involved trade-offs. As an example of this approach, the design space of the PA designed in [11] was plotted in the Fig. 2(b).

This paper presents the synthesis of the analytic relationship between the DC input voltage and V_{Sm}, on a class-E PA with finite dc-feed inductance used in [11].

II. IDEAL NOMINAL CLASS-E MODEL

The ideal class-E PA circuit shown in Fig.1(a) can be modeled as the circuit shown in Fig. 1(b) when the control

![Image](a) Ideal model

![Image](b) Model High Q

Fig. 1. Class-E PA

![Image](a) Gains

![Image](b) Design Space

Fig. 2. Ideal class-E design set.
\[ 0 = \cos (q \theta_m) \sin (2\pi q) \cos (\varphi) pq^2 - \cos (2\pi q) \sin (q \theta_m) \cos (\varphi) pq^2 - \sin (2\pi q) \sin (q \theta_m) \sin (\varphi) pq + \cos (q \theta_m) \sin (2\pi q) q^2 \]

\[ + pq \sin (\theta_m + \varphi) - \cos (q \theta_m) \sin (2\pi q) + \cos (2\pi q) \sin (q \theta_m) - \cos (q \theta_m) \cos (2\pi q) \sin (\varphi) pq - \cos (2\pi q) \sin (q \theta_m) q^2 \]

(14)

signals has 0 time transitions at a frequency \( f \) (near to \( f_0 \)) and
the series resonant circuit \( L_0, C_e \) and \( R_L \) has a high loaded
quality factor. Therefore, the output current is given by (1). As
a result of development presented presented in [5], for ideal
class-E nominal operation (i.e. with ZVS and DZVS), the PA
currents and PA voltages, are given by the equations (2) to (7).

\[ i_R(t) = I_P \sin(\omega t + \varphi) = \sqrt{\frac{2P_{OUT}}{R_L}} \sin (2\pi ft + \varphi); \quad (1) \]

\[ v_{C_{SH,ON}}(t) = i_{C_{SH,ON}}(t) = i_{S_{OFF}}(t) = 0; \quad (2) \]

\[ v_{C_{SH,OFF}}(t) = V_{CC} - I_P \sin(\varphi); \quad (3) \]

\[ i_{S_{ON}}(t) = \frac{V_{CC}}{L_{SH}} t + I_P (\sin(\omega t + \varphi) - \sin(\varphi)); \quad (4) \]

\[ v_{C_{SH,OFF}}(t) = V_{CC} + C_1 \cos(q \omega t) + C_2 \sin(q \omega t) \]

\[ - \frac{q^2}{2} \frac{P_{OUT}}{V_{CC}} \cos(\omega t + \varphi); \quad (5) \]

\[ i_{C_{SH,OFF}}(t) = \frac{V_{CC}}{L_{SH}} t - \frac{1}{L_{SH}} \int_{\frac{2\pi}{\omega}}^{\frac{2\pi}{\omega}} v_{C_{SH}}(\tau)d\tau; \quad (6) \]

\[ + I_P (\sin(\omega t + \varphi) - \sin(\varphi)) \]

\[ i_{L_{SH,OFF}}(t) = \frac{V_{CC}}{L_{SH}} t - \frac{1}{L_{SH}} \int_{\frac{2\pi}{\omega}}^{\frac{2\pi}{\omega}} v_{C_{SH}}(\tau)d\tau - I_P \sin(\varphi); \quad (7) \]

where, \( X_{ON} \) means that the expression \( X \) is valid when the
switch is in the ON state \( (0 < t < \frac{2\pi}{\omega}) \). On the other hand,
\( X_{OFF} \) means that the expression \( X \) is valid when the switch is
in the OFF state \( (\frac{2\pi}{\omega} < t < \frac{4\pi}{\omega}) \). The constants \( C_1 \) and \( C_2 \)
are analytic functions of \( p, q, \varphi \) and \( V_{DD} \), and they were found
in [5]. The variables \( p \) and \( q \) were introduced by [5], in order
to simplify the math analysis and are defined as:

\[ q = \frac{1}{\omega \sqrt{L_{SH} C_{SH}}} = \frac{\omega}{\omega}; \quad p = \frac{\omega}{V_{CC}} \frac{L_{SH} I_P}{R_\omega} = \frac{Z_{LSH}}{R_\omega}; \quad (8) \]

where, \( \omega \) is the parallel tuned frequency of the \( LC_{SH} \)

\[ R_\omega = \frac{V_{CC}}{I_P} = \sqrt{\frac{P_{OUT}}{2R_L}} = \frac{\sqrt{R_L R_{DC}}}{2}; \quad (9) \]

where, \( R_L \) is the PA load, \( P_{in} \) is the power delivered by \( V_{CC} \),
\( P_{out} \) is the power dissipated by \( R_L \), and \( R_{DC} \) is the equivalent
resistance (introduced by [11]) that the amplifier imposes to \( V_{CC} \) and
can be calculated as:

\[ R_{DC} = \frac{V_{DC}}{I_{DC}} = V_{CC} \int_0^{2\pi} \frac{1}{2\pi} i_x(t) dt = \frac{R_\omega}{q}; \quad (10) \]

therefore, from (10) and (9), the \( R_{DC} \) can be rewritten as:

\[ R_{DC} = R_L/2g^2 \quad (11) \]

and is the impedance of \( L_{SH} \) and \( R_\omega \) is:

\[ \frac{d}{dt} V_C(\omega \tau_{max}) = 0; \quad (13) \]

where, \( \tau_{max} \) is the time at which the peak value occurs. From
(5) and (13) it was found (14) (i.e. the equation at the top of
the page). Solving numerically this equation, the \( \omega \tau_{max} \) value
is calculated in function of the parameter values \( q \) and \( D \),
using \( \omega \tau_{max} \) the peak value of the switch voltage value gain
\( G_V(q, D) = \frac{V_{max}}{V_{max}} \) was found from (6), and the results were
plotted in the Fig.3.

### III. IDEAL CLASS-E PA SWITCH BREAKDOWN VOLTAGE

#### A. Numerical solution

The peak value of the switch voltage occurs when:

\[ \frac{d}{dt} V_C(\omega \tau_{max}) = 0; \quad (13) \]

where, \( \tau_{max} \) is the time at which the peak value occurs. From
(5) and (13) it was found (14) (i.e. the equation at the top of
the page). Solving numerically this equation, the \( \omega \tau_{max} \) value
is calculated in function of the parameter values \( q \) and \( D \),
using \( \omega \tau_{max} \) the peak value of the switch voltage value gain
\( G_V(q, D) = \frac{V_{max}}{V_{max}} \) was found from (6), and the results were
plotted in the Fig.3.

#### B. Curve fitting

As a simplified approach, the variation on \( G_V \) gain with
respect to changes in the \( q \) variable can be neglected, therefore
this gain was assumed as:

\[ G_V(a, D, q) = \frac{a}{1 - D}; \quad (15) \]

where, \( a \) is a constant value that minimizes the involved error.
The fitting error was defined following the percentage least
squares criteria as:

\[ E_k = \frac{(G_V(a, q, D_j) - G_V(q, D_j))}{G_V(q, D_j)} \quad (16) \]
where \( q_i \) is the sample \( i \) of the \( q \) variable and \( q_j \) is the sample \( j \) of the \( D \) variable, and \( E_k \) is the error value for the sample \((i, j)\). The goal function \((S)\) was defined as:

\[
S(a) = \frac{\sum_k E_k^2}{n} \quad (17)
\]

where, \( n \) is the number of samples of the numerical solution. Therefore, the optimum value of \( a \) minimizes the sum of the calculated error squares for all the samples. The optimization process is illustrated in the Fig. 4(b). The error function for this optimum value was plotted in Fig.4(a). Using the optimum value, (15) can be rewritten as:

\[
G_V a(D, q) = \frac{1.8551}{1 - D}. \quad (18)
\]

In order to analyze this results, the absolute value of the percentage error was plotted in Fig.5. Considering this figure, the approximation error is less than 5% for any \( D \) value and a \( q \) less than 1.7. For large \( q \) (i.e. \( q > 1.65 \)) the values of either input parameters or circuit elements correspond to extreme values [6]. Hence, we focus on \( q \) values lower than 1.7. The mean perceptual error for the limited range \((0 < q < 1.7, 0 < D < 1)\) was 3%, and the maximum perceptual error was 7%. Further, a similar optimization procedure was made. Therefore, (15) can be rewritten as:

\[
G_V a(D, q) = \frac{1.8208}{1 - D}; \quad (19)
\]

the absolute value of the new percentage error was plotted in Fig.6. Considering this figure, the approximation error is less than 5.3% for any \( D \) value and a \( q \) less than 1.5. The mean perceptual error in all the range \((0 < q < 1.7, 0 < D < 1)\) was 2%, and the maximum perceptual error was 8.8%.

In order to reduce the approximation error in the peak value prediction, for a fixed duty cycle \((D = D_x)\), the approximative expression may be refined using a polynomial \( c(x) \) of degree \( n \) of the variable \( q \) that fits the data, in the least squares sense. Therefore (15) can be rewritten as:

\[
G_V a(D_x, q) = \frac{c(q)}{1 - D_x} = \frac{c_0 + c_1 q + \cdots + c_n q^n}{1 - D_x}; \quad (20)
\]

using a duty cycle of 50%, the error can be decreased as is presented in the Fig.7. Hence, using \( n = 2 \), (20) can be rewritten as:

\[
G_V a(D_x, q) = \frac{1.7613 + 0.0500q}{1 - D_x}; \quad (21)
\]

this expression has perceptual error of less than 2% in all the range \((0 < q < 2, D = 50\%\)

IV. USING THE CLASS-E DESIGN SET

The relations between the input parameters \((Q_L, V_{CC}, P_{OUT} \text{ and } \omega)\) and the circuit element values \((L_S, C_S, C_e, L_0 \text{ and } R_L)\) should be known. These relations are commonly referenced as the design set, in this paper we use the set proposed in [5] and extended in [11]. This set is summarized in the table I.

| \( K_P(q, D, V_{CC}, R_L) = \frac{P_{OUT}}{R_L} = 2g(q, D)^2 \left( \frac{V_{CC}}{R_L} \right)^2 \) |
| \( K_C^{-1}(q, D, \omega) = (C_{SH})^{-1}/R_L = \frac{2\pi g(q, D)}{\omega} \) |
| \( K_L(q, D, \omega) = L_S/H/R_L = \frac{\mu(q, D)}{\omega^2 + \frac{1}{\omega^2} + \frac{1}{\omega^2}} \) |
| \( K_X = (q, D) = \frac{X_S}{R_L} = \frac{v_{CC}/q}{C_{Q}^{1}/R_Q} = \frac{\int_{-\frac{1}{2\omega}}^{\frac{1}{2\omega}} \frac{\int_{-\frac{1}{2\omega}}^{\frac{1}{2\omega}} e^{(t+k)(\omega t+\omega)} dt}{\int_{-\frac{1}{2\omega}}^{\frac{1}{2\omega}} e^{(t+k)(\omega t+\omega)} dt} dt}{\int_{-\frac{1}{2\omega}}^{\frac{1}{2\omega}} e^{(t+k)(\omega t+\omega)} dt} \) |
| \( K_{L_{-1}}(q, Q_L) = L_0/R_L = \frac{Q_L}{L_0} \) |
| \( K_{C_{-1}}(q, Q_L) = (C_0)^{-1}/R_L = Q_L \omega \) |
| \( K_{C_{-1}}(\omega, Q_L, q, D) = (C_0)^{-1}/R_L = \omega (Q_L - K_X(q, D)) \) |

V. MEASUREMENT AND SIMULATED RESULTS

In order to verify (19), (21) and (11), a PA was simulated following the specifications summarized in Table II. The simulation setup uses the harmonic balance simulation technique in the Advanced Design System (ADS®) software. Furthermore, the transistor is represented as a voltage controlled switch model, with ideal control signal (with 0 time transitions at a frequency \( f \)), an on resistance of 1 mΩ, and an open resistance of 100 GΩ. All the other circuit elements are simulated as ideal components. In order to calculate its values, first the relations that are only dependent on the \( q \) and \( D \) specification were
calculated (summarized in the Table III), then following the circuit element gains of the design set and the $Q_L$ specification, the circuit values were calculated and summarized in Table IV.

The experimental results are taken from [6], where discrete Class-E PAs were constructed with a transistor (MAX 2601), and discrete passive components. Further, the transistor gate was driven by a square wave signal from the signal source.

Table VI shows the calculated circuit values.

### Table III: Relations that are only functions of $q$ and $D$.

<table>
<thead>
<tr>
<th>$D$ (%)</th>
<th>$q$</th>
<th>$p$</th>
<th>$w$</th>
<th>$g$</th>
<th>$K_p$</th>
<th>$K_q$</th>
<th>$K_{q,D}$</th>
<th>$R_D$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.8</td>
<td>0.8</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
</tr>
</tbody>
</table>

### Table IV: Calculated circuit values.

<table>
<thead>
<tr>
<th>$L_0$ (μH)</th>
<th>$C_0$ (pF)</th>
<th>$C_n$ (μF)</th>
<th>$L_{SH}$ (μH)</th>
<th>$C_{SH}$ (pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Q=6</td>
<td>2.052</td>
<td>117.746</td>
<td>138.271</td>
<td>117.743</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>102.659</td>
<td>1975</td>
</tr>
<tr>
<td>Q=10</td>
<td>3.419</td>
<td>70.648</td>
<td>77.555</td>
<td>70.646</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>64.923</td>
<td>250.691</td>
</tr>
<tr>
<td>Q=100</td>
<td>34.193</td>
<td>7.065</td>
<td>7.128</td>
<td>7.005</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>7.005</td>
<td>253.066</td>
</tr>
</tbody>
</table>

### Table V: Class-E PA results.

<table>
<thead>
<tr>
<th>$P_{out}$ (mW)</th>
<th>$P_{DC}$ (mW)</th>
<th>$V_{SM}$ (Num, Eq19) (V)</th>
<th>$R_{DC}$ (Ω)</th>
<th>$V_{SM}$ (Num, Eq21) (V)</th>
<th>$R_{DC}$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>136.8</td>
<td>138.0</td>
<td>140.6</td>
<td>150</td>
<td>134.2</td>
<td>145.0</td>
</tr>
<tr>
<td>136.8</td>
<td>138.0</td>
<td>141.0</td>
<td>149</td>
<td>134.2</td>
<td>145.0</td>
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<tr>
<td>136.8</td>
<td>140.4</td>
<td>141.0</td>
<td>149</td>
<td>134.2</td>
<td>145.0</td>
</tr>
<tr>
<td>136.8</td>
<td>140.4</td>
<td>141.0</td>
<td>149</td>
<td>134.2</td>
<td>145.0</td>
</tr>
</tbody>
</table>

### Table VI: Error results.

<table>
<thead>
<tr>
<th>$P_{out}$ (mW)</th>
<th>$P_{DC}$ (mW)</th>
<th>$V_{SM}$ (Num, Eq19) (V)</th>
<th>$R_{DC}$ (Ω)</th>
<th>$V_{SM}$ (Num, Eq21) (V)</th>
<th>$R_{DC}$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>136.8</td>
<td>138.0</td>
<td>140.6</td>
<td>150</td>
<td>134.2</td>
<td>145.0</td>
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<td>136.8</td>
<td>138.0</td>
<td>141.0</td>
<td>149</td>
<td>134.2</td>
<td>145.0</td>
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<tr>
<td>136.8</td>
<td>140.4</td>
<td>141.0</td>
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<td>134.2</td>
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<tr>
<td>136.8</td>
<td>140.4</td>
<td>141.0</td>
<td>149</td>
<td>134.2</td>
<td>145.0</td>
</tr>
</tbody>
</table>

### VI. Conclusions

An analytical expression of the gain between the DC input voltage and the peak switch voltage on a ideal class-E power amplifier (PA) for a finite dc-feed inductance and ZVS and DZVS operation was presented. This expression was verified by the simulations, and was evaluated by experimental results ($f = 10.24$ MHz), with good agreement between the results and the predicted values. Considering the simulated and experimental results the maximum predicted error was $10.2\%$.

### Acknowledgment

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### References


