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SUBMISSION DATE / POSTED DATE

25-08-2021 / 27-08-2021

CITATION

yao, jinbo (2021): Analysis and Design of Class E Frequency Multiplier With Shunt Linear and Nonlinear Capacitances at Any Duty Cycle. TechRxiv. Preprint. <https://doi.org/10.36227/techrxiv.16436178.v1>

DOI

[10.36227/techrxiv.16436178.v1](https://doi.org/10.36227/techrxiv.16436178.v1)

Analysis and Design of Class E Frequency Multiplier With Shunt Linear and Nonlinear Capacitances at Any Duty Cycle

Jinbo Yao, Mingyu Li, *Member, IEEE*, Xuefeng Yu, Tian Li, and Zhijiang Dai, *Member, IEEE*,

Abstract—The theoretical efficiency of the class E frequency multiplier with arbitrary output frequency can reach 100% by controlling the phase of the output current to satisfy the ZVS/ZVDS condition. It is necessary to select the appropriate value of the shunt capacitance of the frequency multiplier for satisfying these two conditions. In general, the shunt capacitance of the transistor consists of the linear external capacitance and the nonlinear parasitic capacitance. Therefore, the conventional design method of treating shunt capacitance as purely linear is not accurate enough. In this paper, a class E frequency multiplier with shunt linear and nonlinear capacitances at any duty cycle is analyzed and designed, and the design curve is given in detail. The simulation and experimental results shows highly agreement with the theoretical analysis result.

Index Terms—Class E, frequency multiplier, zero-voltage switching (ZVS), zero-voltage-derivate switching(ZVDS)

I. INTRODUCTION

THE class-E power amplifiers (PAs) have been concerned by the academic community for their simple circuits, and the extremely higher efficiency can be obtained when the zero-voltage switching and zero-voltage derivative switching (ZVS/ZVDS) conditions are satisfied [1]-[4]. The Class E frequency multiplier, which only changes the resonant frequency of the series resonant circuit on the Class E circuit, also has the above two characteristics and can be used as the driver circuit [5]. And the mathematical derivation and design equation for the frequency multiplier are analyzed and given in [6]-[8]. These papers indicate that the duty cycle of the frequency multiplier must be less than $1/n$ (n refers to the multiple of the frequency multiplier output frequency to the drive signal), otherwise the drain-source voltage will be negative and is harmful to the circuit. Furthermore, the influence of duty cycle on the performance of frequency multiplier is analyzed in [9] and design equations under different Q and n values are given in detail. Meanwhile, two different topologies to enable the frequency multiplier to work with duty cycle of 0.5 are proposed in [10]-[11]. However, the shunt capacitors which are composed of linear capacitors and nonlinear capacitors are not take into account in these papers.

Manuscript received xx, 2021; revised xx, 2021; accepted xx, 2021.(Corresponding Author: Mingyu Li)

This work was supported by the National Natural Science Foundation of China under 62171068.

Jinbo Yao, Mingyu Li, Tian Li, and Zhijiang Dai are with School of Microelectronics and Communication Engineering, Chongqing University, Chongqing 400044, China(e-mail: myli@cqu.edu.cn). Xuefeng Yu is with the Xi'an branch of China Academy of Space Technology, Xi'an 710100, China.

In fact, the shunt capacitance value is vital for achieving the optimum conditions for the class E circuit. Some papers have been pointed out that the linearity and nonlinearity of the shunt capacitance have different effects on the performance of the class E circuit [12]-[17]. For example, the case that the shunt capacitance is composed of linear capacitance and nonlinear capacitance is considered for the class E amplifier design in [16], and the design equation through numerical solution is given only when the duty cycle is 0.5. Furthermore, the design of the class E amplifier with the shunt linear and nonlinear capacitances under arbitrary duty cycle is proposed in [17]. In the practical application of the frequency multiplier circuit, the inherent nonlinear parasitic capacitance of the MOSFET is difficult to be exactly applicable for the design specification. And a suitable external capacitor is usually required to make the total shunt capacitance conforming to the design specification. Therefore, the effects of the shunt linear and nonlinear capacitances on the performance of the frequency multiplier circuit must be taken into account.

In this paper, the analytical derivation of the class E frequency multipliers with shunt linear and nonlinear capacitance, which satisfies the ZVS/ZVDS conditions at any duty ratio, is given. The mathematical relationship of the circuit parameters as the functions of the duty ratio are derived in detail. A frequency doubler circuit example is designed for verifying the proposed design method. The simulation and the experimental results show the usefulness and effectiveness of the analytical design approach for the class E frequency multipliers at any duty ratio.

II. CIRCUIT ANALYSIS

A basic circuit of class E frequency multiplier is shown in Fig.1. In this circuit, C_o is the device's device's parasitic nonlinear capacitance, and C_e is the external linear capacitance, C_o and C_e make up the shunt capacitance of the frequency multiplier circuit together. The analysis of the class E frequency multiplier of Fig.1 is carried out under the following assumptions.

- 1) The inductance of the choke coil L_{RFC} is high enough that the current I_{DD} flowing through it could be considered constant.
- 2) The loaded quality factor Q of the circuit is large enough that the output current waveform is pure sinusoidal.
- 3) All component resistances and switch-on resistances are ignored, so that the input power is equal to the output power if the soft switch condition is satisfied.

Fig. 1. Basic circuit of class E multiplier

- 4) The device switch turns on and turns off so quick, that it can be considered instantaneous and no power loss.
- 5) The duty cycle is arbitrary under the basic limits of the frequency multiplier.
- 6) The total shunt capacitance C_{ds} is equal to the nonlinear capacitance C_o plus the linear capacitance C_e , which is described by

$$C_{ds} = C_o + C_e = \frac{C_{j0}}{\left(1 + \frac{V_S}{V_{bi}}\right)^m} + C_e \quad (1)$$

where C_{j0} is the output capacitance of the device when the drain-to-source voltage $V_s = 0$, V_{bi} is the built-in potential of the MOSFET body diode, the grading coefficient m is a parameter determined by the device, which is defined as $m = 0.5$ in this paper. The nonlinearity of the shunt capacitance is incorporated into the analysis through this equation.

Defining the state of the switch as off when $0 < \theta \leq 2\pi D$ (D is the duty cycle), so no current flows through the switch, i.e. $I_S = 0$, in this period, applying the KCL law, the current on the overall shunt capacitance is expressed as

$$I_C(\theta) = I_{DD} - I_n \sin(n\theta + \Phi) \quad (2)$$

where I_{DD} is the dc input current, I_n is the amplitude of the n th harmonic output current, Φ is the phase shift of the output current and $\theta = \omega t$. Considering the voltage and current relationship across the capacitance, the current through the shunt capacitance could be redefined as

$$I_C(\theta) = I_{C_o}(\theta) + I_{C_e}(\theta) = \omega(C_o + C_e) \frac{dV_S}{d\theta} \quad (3)$$

where $I_{C_o}(\theta)$ represents the current flowing through the nonlinear capacitance C_o , $I_{C_e}(\theta)$ represents the current flowing through the linear capacitance C_e , V_S represents the voltage across the shunt capacitance. Integrating both sides of this equation with respect of θ , the following expressions can be given

$$\int I_C(\theta) d\theta = \omega \int \left(\frac{C_{j0}}{\sqrt{1 + \frac{V_S}{V_{bi}}}} + C_e \right) dV_S \quad (4)$$

Then substituting (2) into (4), and expanding the integral of both sides, a quadratic equation of V_S is yielded. There are two

Fig. 2. Waveforms of the switch voltage for $g = \infty, g = 2.5$ and $g = 0$ at $V_{DD}/V_{bi} = 14.29$, $D = 0.25$, $n = 2$.

solutions for this equation, and one of the solutions can not satisfy the ZVS condition, which is dropped. Another solution can satisfy the ZVS condition, which is given as

$$V_S = 2gV_{bi}(g + h(\theta) - \sqrt{g^2 + 2gh(\theta) + 1}) \quad (5)$$

where

$$h(\theta) = \frac{I_{DD}\theta + \frac{I_n}{n}(\cos(n\theta + \Phi) - \cos(\Phi))}{2\omega C_{j0}V_{bi}} + 1 \quad (6)$$

$$g = \frac{C_{j0}}{C_e} \quad (7)$$

where $h(\theta)$ is an intermediate form for simplifying the equation. And g is the ratio of C_{j0} to C_e , which represents the degree of nonlinearity in the shunt capacitance. Fig.2 shows the effect of shunt capacitance on the V_S waveform of a frequency doubler at $D = 0.25$ in three typical cases, which satisfies the optimum conditions. The three cases are: $g = \infty, g = 2.5$ and $g = 0$ at $V_{DD}/V_{bi} = 14.29$. The derivation of V_S at $g = \infty$ is obtained from [12], the waveform at $g = 2.5$ is obtained from (5), the waveform at $g = 0$ is obtained in [7]. Here $g = \infty$ implies that the shunt capacitance entirely consists of nonlinear capacitance, and the peak voltage in this case is significantly higher than the other two cases. And $g = 2.5$ implies the shunt capacitance consists of both nonlinear and linear capacitance, which represents the most common situation in practical applications. The peak voltage in this case is between the other two cases. Finally, $g = 0$ implies the shunt capacitance only consist of linear capacitance. The peak voltage is the lowest in three cases.

III. CIRCUIT PARAMETERS FOR OPTIMUM OPERATION AT ANY DUTY RATIO

A. ZVS Condition

Applying the ZVS condition, that is

$$V_S(2\pi D) = 0 \quad (8)$$

Substituting (8) into (5)

$$h(2\pi D) = \frac{2\pi D I_{DD} + \frac{I_n}{n}(\cos(2n\pi D + \Phi) - \cos(\Phi))}{2\omega C_{j0}V_{bi}} + 1 = 1 \quad (9)$$

Fig. 3. $\omega C_{j0}R$ versus D for selected values of g at $V_{DD}/V_{bi} = 14.29$, $n = 2$.

The relationship between I_{DD} and I_n can be revealed

$$I_{DD} = I_n \frac{\cos \Phi - \cos(x + \Phi)}{x} \quad (10)$$

Where $x = 2\pi D$, and ignoring power losses, input power equals output power, then

$$I_{DD}V_{DD} = I_n^2 R / 2 \quad (11)$$

combining (10) and (11), the amplitude of the output current can be expressed by V_{DD} and R

$$I_n = \frac{2V_{DD}}{R} \frac{\cos \Phi - \cos(x + \Phi)}{x} \quad (12)$$

And the amplitude of output voltage is

$$V_n = RI_n = 2V_{DD} \frac{\cos \Phi - \cos(x + \Phi)}{x} \quad (13)$$

The dc input current is

$$I_{DD} = \frac{I_n^2 R}{2V_{DD}} = \frac{2V_{DD}}{R} \left(\frac{\cos \Phi - \cos(x + \Phi)}{x} \right)^2 \quad (14)$$

The output power can be expressed as

$$P_o = \frac{I_n^2 R}{2} = \frac{2V_{DD}}{RV_{DD}} \left(\frac{\cos \Phi - \cos(x + \Phi)}{x} \right)^2 \quad (15)$$

B. ZVDS Condition

Introducing the ZVDS condition, that is

$$I_C(2\pi D) = 0 \quad (16)$$

Substituting (16) into (2)

$$I_{DD} = I_n \sin(x + \Phi) \quad (17)$$

Then uniting (10), the following expression can be obtained

$$\tan \Phi = \frac{\cos x + x \sin x - 1}{\sin x - x \cos x} \quad (18)$$

$$\sin \Phi = \frac{\cos x + x \sin x - 1}{Z} \quad (19)$$

$$\cos \Phi = \frac{\sin x - x \cos x}{Z} \quad (20)$$

where

$$Z = \sqrt{x^2 + 2 - 2 \cos x - 2x \sin x} \quad (21)$$

Substituting (19) and (20) into (12)-(15), the following expressions can be obtained

$$I_n = \frac{2V_{DD}}{R} \frac{1 - \cos x}{Z} \quad (22)$$

$$V_n = 2V_{DD} \frac{1 - \cos x}{Z} \quad (23)$$

$$I_{DD} = \frac{2V_{DD}}{R} \frac{(1 - \cos x)^2}{Z^2} \quad (24)$$

$$P_o = \frac{2V_{DD}^2}{R} \frac{(1 - \cos x)^2}{Z^2} \quad (25)$$

C. Component Values

Refer to the assumptions 3), there is no resistance on L_{RFC} , which means

$$\begin{aligned} V_{DD} &= \frac{1}{2\pi} \int_0^{2\pi} V_S d\theta \\ &= \frac{1}{2\pi} \int_0^{2\pi D} 2gV_{bi}(g + h(\theta) - \sqrt{g^2 + 2gh(\theta) + 1}) d\theta \end{aligned} \quad (26)$$

Then (26) can only be partially expanded as

$$\begin{aligned} \frac{V_{dd}}{V_{bi}} &= \frac{g}{\pi} (2\pi D (g + 1) + \frac{2V_{DD}}{R} \frac{1 - \cos x}{Z} (\sin(x + \Phi) \\ &\quad - \sin \Phi + \frac{x^2(1 - \cos x)}{2Z} - x \cos \Phi) / (2\omega C_{j0} V_{bi} n^2) \\ &\quad - \int_0^{2\pi D} \sqrt{g^2 + 2gh(\theta) + 1} d\theta) \end{aligned} \quad (27)$$

As can be seen in (27), $\omega C_{j0}R$ is a function of g and D . Fig.3 shows the variation relationship of $\omega C_{j0}R$ with respect to D when $V_{DD}/V_{bi} = 14.29$ for selected g . The maximum value of D is 0.5 when $n = 2$. Setting g value from 0 to infinity, it can be seen that $\omega C_{j0}R$ is always 0 when $g = 0$ because there is no nonlinear capacitance in the shunt capacitance, which means $C_{j0} = 0$. when the value of g grows, then the $\omega C_{j0}R$ value gets bigger and bigger, and $\omega C_{j0}R$ reaches the maximum value when $g = \infty$, which means that the shunt capaciancer only contains nonlinear capacitance. On the other hand, the $\omega C_{j0}R$ value value increases from 0 to its peak as D increases from 0 to 0.324 and then decrease to 0 as $D = 0.5$ for all selected g values. The n th harmonic component $V_{Sn}(\theta)$ of the switching voltage V_S contains the output voltage $V_o(\theta)$ and the voltage $V_{Lx}(\theta)$ across the reactance L_x

$$\begin{aligned} V_{Sn}(\theta) &= V_o(\theta) + V_{Lx}(\theta) \\ &= V_o \sin(n\theta + \Phi) + \frac{X}{R} V_o \cos(n\theta + \Phi) \\ &= V_n \sin(n\theta + \Phi_1) \end{aligned} \quad (28)$$

where $X = n\omega L_x$. The amplitude V_n can be expressed as

$$V_n = V_o \sqrt{1 + \left(\frac{X}{R}\right)^2} \quad (29)$$

Fig. 4. V_{SM} versus D for selected $\omega C_{j0}R$, at $V_{DD}/V_{bi} = 14.29$, $n = 2$, $Q = 15$.

Fig. 5. C_p versus $\omega C_{j0}R$ for selected D at $V_{DD}/V_{bi} = 14.29$, $n = 2$

and the phase of Φ_1 can be determined as

$$\Phi_1 = \Phi + \arctan\left(\frac{X}{R}\right) \quad (30)$$

The n th order fourier expansion of the switching voltage V_S is expressed as

$$\frac{1}{\pi} \int_0^{2\pi D} V_S \cos(n\theta + \Phi_1) d\theta = 0 \quad (31)$$

Transforming this expression, then Φ_1 is described as

$$\tan \Phi_1 = \frac{\int_0^{2\pi D} (g + h(\theta) - \sqrt{g^2 + 2gh(\theta) + 1}) \cos n\theta d\theta}{\int_0^{2\pi D} (g + h(\theta) - \sqrt{g^2 + 2gh(\theta) + 1}) \sin n\theta d\theta} \quad (32)$$

Then X is obtained

$$X = R \tan(\Phi_1 - \Phi) \quad (33)$$

Once X is given, the redundant inductance L_x can be obtained

$$L_x = R \tan(\Phi_1 - \Phi) / (n\omega) \quad (34)$$

The inductance L can be derived

$$L = \frac{RQ}{n\omega} \quad (35)$$

and the series capacitance C is given by

$$C = \frac{1}{(n\omega)^2 (L - L_x)} \quad (36)$$

D. Output Power Capability

The switch peak current I_{SM} is the sum of the dc supply current and the amplitude of the output current I_n

$$\begin{aligned} I_{SM} &= I_{DD} + I_n \\ &= \frac{2V_{DD}}{R} \frac{1 - \cos x}{Z} \left(1 + \frac{1 - \cos x}{Z}\right) \end{aligned} \quad (37)$$

The switch peak voltage V_{SM} occurs when the shunt current I_C is zero at $\theta = \theta_1$, i.e., $V_{SM} = V_{S(\theta_1)}$, where

$$\theta_1 = \frac{-\Phi + \arcsin((1 - \cos x)/Z)}{n} \quad (38)$$

Fig. 6. photograph of the experimental frequency doubler

Fig.4 shows the relationship between the peak voltage V_{SM} and D for different values of $\omega C_{j0}R$. The following conclusions can be obtained from the figure. First, the value of $\omega C_{j0}R$ affects the value range of D when the peak voltage V_{SM} exists. The larger the value of $\omega C_{j0}R$, the smaller the range of the values of D . Then, the peak voltage V_{SM} shows a downward trend, and the downward trend becomes slower with the increase of D . Finally, the value of V_{SM} is slightly higher if $\omega C_{j0}R$ takes a larger value for the same D value. The output power capability C_p is defined as

$$C_p = \frac{I_{DD}V_{DD}}{I_{SM}V_{SM}} \quad (39)$$

Fig.5 shows the relationship between C_p and D for selected $\omega C_{j0}R$ at $V_{DD}/V_{bi} = 14.29$, $n = 2$. It can be seen that the selection of $\omega C_{j0}R$ determines the range of D in which C_p exists. C_p is slightly higher when the value of $\omega C_{j0}R$ is smaller. And the value of D at the peak of C_p increases with the increment of $\omega C_{j0}R$.

IV. SIMULATION AND EXPERIMENTAL RESULTS

An example frequency doubler circuit is given to verify this theory and its design specifications are $V_{DD} = 10V$, $R = 20\Omega$, $D = 0.25$ and $Q = 15.25$ at operating frequency $f = 1.5MHz$. In this experiment, The transistor IRF510 was selected, and the parameter value is $C_{j0} = 400pF$, $V_{bi} = 0.7V$. The fabricated experimental frequency doubler is shown in Fig. 6. The value of g can be determined as

TABLE I
THEORETICAL AND SIMULATED PERFORMANCE
CLASS E DOUBLER, $Q = 15.25$

	Theory Values	Simulation Values	Experiment Values	Difference %
R, Ω	20	20	20	0
$L, \mu H$	16.18	16.2	16.18	0
C, pF	188.69	188.3	196	3.8%
C_e, pF	159.69	164	140	-12.3%
V_{SM}, V	75.32	79.32	73	-2.9%
I_{SM}, A	0.8254	0.875		
C_p	0.0463	0.0438		
$V_o(rms), V$	7.6	7.65	7.33	-3.5%
P_{in}, W	2.884	3.04	2.85	-1.2%
P_{out}, W	2.884	2.924	2.686	-3.7%
$\eta, \%$	100	96.18	94.2	-5.8%
D	0.25	0.25	0.25	0%

Fig. 7. The simulated and experimental waveform of V_S and V_o

2.5 from (27). Then the values of all the circuit components can then be computed, the values of components in the circuit and some important circuit performance parameters was given in Table I. The series resonate inductor L was made in our lab and the value was measured by KEYSIGHT E5063A. As can be seen in Table I, the theoretical value, simulation value and experimental value of each component are very consistent. Fig.7 shows the drain-to-source voltage and output voltage waveforms obtained from ADS simulation and experimental results. It can be seen from Fig.7 and Table I that the theory, simulation and experimental results agrees well, which strongly proves the validity of the theory proposed in this paper.

V. CONCLUSION

An analysis, simulation and experiment of class E frequency multiplier with shunt nonlinear and linear capacitances at any duty ratio was presented. The effect of the duty ratio frequency multiplier is analyzed under the premise of the shunt capacitances composed of nonlinear and linear capacitances. A design equation for any harmonic output class E frequency multiplier was given. The theoretical results was validated by circuit simulation and experiment. The calculated and measured results were in great agreement.

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