

A small-sized class-J power amplifier from combined multi-harmonic voltage reflection functions

Youngcheol Park^{a)} and Jongkyun Kim

*Hankuk University of Foreign Studies,
Mohyun-meon, Yongin-si, Kyunggi-do 449-791, South Korea*

a) ycpark@hufs.ac.kr

Abstract: In this study, multi-harmonic voltage reflections are analyzed for compact power amplifiers that can be integrated into micro-sized radio units. By combining impedances at DC and harmonic frequencies throughout the transmission lines of power amplifier circuits, this method is used to reduce analytically the footprint of these circuits. Further, a cost function that takes the occupied area and the circuit performance into account is used to measure the effectiveness of the suggested method. For the sake of verification, a 10 W class-J power amplifier is designed for 2.0–2.5 GHz, and showed a 35% reduction in its footprint, a peak efficiency of 71.3%, an output power of 40 dBm.

Keywords: harmonics, power amplifier

Classification: Microwave and millimeter wave devices, circuits, and systems

References

- [1] Alcatel-Lucent, *Technology White Paper: lightRadio Portfolio – Technical overview*, 2011.
- [2] D. M. Pozar, *Microwave Engineering*, Wiley, Hoboken, 1997.
- [3] S. C. Cripps, P. J. Tasker, A. L. Clarke, J. Lees, and J. Benedikt, “On the Continuity of High Efficiency Modes in Linear RF Power Amplifiers,” *IEEE Microw. Wireless Compon. Lett.*, vol. 19, no. 10, pp. 665–667, Oct. 2009.
- [4] Y. Park, “Expansion of Class-J Power Amplifiers into Inverse Mode Operation,” *IEICE Electron. Express*, vol. 8, no. 18, pp. 1479–1485, 2011.
- [5] *Evaluation Board for GH40010 RF Power GaN HEMT*, CREE Inc., Dunham, NC, 2012.

1 Introduction

With the proliferation of mobile devices, the demand for new architectures to deliver high data throughput has escalated. One of the approaches to meet this requirement is to use distributed and reconfigurable base stations, where a micro-sized radio unit is one of the underlying devices [1]. This

paper introduces a systematic approach to reduce the size of power amplifier circuits. By combining the reflected voltage functions at DC and harmonic frequencies, an optimized circuit is found that exhibits a reduced footprint and satisfies the required impedance conditions at designated frequencies. Based on the analysis on voltage equations in transmission at harmonic frequencies, a 10-W class-J power amplifier is implemented at 2.0~2.5 GHz with the conditions applied to the matching network. Then, its performance is evaluated and discussed afterwards.

2 Combined transmission line equations at multiple harmonics

A general circuit for matching networks of wireless amplifiers is shown in Fig. 1, where the DC and AC (fundamental and the second harmonic) paths are terminated with proper impedances. Assuming that the circuit is intended for matching at the input of a transistor, the impedance seen from the right-hand-side port should be zero at the DC and the second harmonic frequencies, and should have a desired value at the fundamental frequency. Generally, ℓ_2 is very close to zero and ℓ_3 is designed to be one-quarter wavelength at the fundamental frequency so that its impedance solely depends on the voltage and the current waves from the source side after the connection of all three impedances. Although there might be some workarounds to reduce the occupied area, such as bending of ℓ_3 , this paper suggests a systematic way to reduce the area by combining the requirements from different frequencies while maintaining its performance. The aforementioned workarounds could then be applied for further size reduction. Looking from the right-hand-side port that connects to the gate of a transistor, we obtain Z_g from the reflection equations throughout the transmission line ℓ_2 [2].

$$Z_g = Z_{o2} \frac{Z_1 || Z_3 + jZ_{o2} \tan \beta \ell_2}{Z_{o2} + jZ_R \tan \beta \ell_2}, \quad (1)$$

where β is the propagation constant, and Z_{o2} is the characteristic impedance of ℓ_2 transmission line.

In general, this circuit has ℓ_3 of one-quarter wavelengths at the fundamental frequency (λ_1), and ℓ_2 is of nearly zero length. The occupied area is then calculated by the circuit span (L), lengths and width (W) of

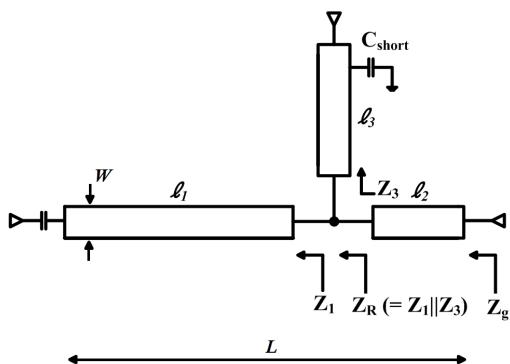


Fig. 1. General circuit for matching with transmission lines that include DC biasing and harmonic termination.

transmission lines, as follows:

$$\text{Area} = (\ell_1 + \ell_2) \times (\ell_3 + W) \approx L \times (\lambda_1/4 + W). \quad (2)$$

When ℓ_3 is reduced by any means, the circuit area is reduced, as well; however, the signal transmission ratio from ℓ_1 through ℓ_2 drops owing to the low impedance in Z_3 from the shunt capacitor connected to ℓ_3 . On the basis of these observations, a figure of merit (FOM) is defined to evaluate the effectiveness of the size reduction while reflecting the signal transmission by the ratio of the impedances. With this measure, lengths of transmission lines and the location of the shunt connection can be found systematically.

$$\begin{aligned} FOM &= \text{Impedance Ratio/Area} \\ &= \left| \frac{Z_3}{Z_1} \right| \frac{1}{(\ell_1 + \ell_2) \times (\ell_3 + W)}. \end{aligned} \quad (3)$$

The following three conditions on Z_g should be met by the input matching network:

Condition A: $Z_g = 0$ at DC

Since $\beta = 0$ at DC, this condition is met unless any series capacitance is incorporated through ℓ_2 and ℓ_3 transmission lines.

$$Z_g = Z_{o2} \frac{Z_R + jZ_{o2} \cdot 0}{Z_{o2} + jZ_R \cdot 0} = Z_1 || Z_3 = Z_3. \quad (4)$$

Condition B: $Z_g = 0$ at $2f_o$

Equation (1) at $2f_o$ is used to obtain the equivalent Z_g when ℓ_3 is terminated into the ground by a capacitor, as follows:

$$\begin{aligned} Z_g &= Z_{o2} \frac{Z_1 || Z_3 + jZ_{o2} \cdot \alpha_2}{Z_{o2} + j(Z_1 || Z_3) \cdot \alpha_2} \\ &= Z_{o2} \frac{-Z_{o2} Z_{o3} \alpha_2 \alpha_3 + jZ_1 (Z_{o2} \alpha_2 + Z_{o3} \alpha_3)}{Z_1 (Z_{o2} - Z_{o3} \alpha_2 \alpha_3) + jZ_{o2} Z_{o3} \alpha_3}, \end{aligned} \quad (5)$$

where α_1 through α_3 represent $\tan(\beta \ell_1)$ through $\tan(\beta \ell_3)$, Z_{o1} and Z_{o3} are characteristic impedances of ℓ_1 and ℓ_3 transmission lines, respectively.

If we assume that Z_1 is greater than Z_3 at the second harmonic, Z_1 can be neglected in Z_R , and the above equation can be made to be equal to zero by the following condition:

$$\begin{aligned} Z_g &= Z_{o2} \frac{j(Z_{o2} \alpha_2 + Z_{o3} \alpha_3)}{Z_{o2} - Z_{o3} \alpha_2 \alpha_3} = Z_{o2} \frac{j(\alpha_2 + \alpha_3')}{1 - \alpha_2 \alpha_3'}, \\ &= jZ_{o2} \tan \beta (\ell_2 + \ell_3') = 0 \end{aligned} \quad (6)$$

where ℓ_3' is the effective length of ℓ_3 ($= \alpha_3 Z_{o3} / Z_{o2}$), and $\alpha_3' = \tan(\beta \ell_3')$.

From the preceding equation, *Condition B* can be satisfied when the effective length of ℓ_2 and ℓ_3 constitutes one-half of the wavelength at the second harmonic.

Condition C: $Z_g = R_g + jX_g$ at f_o

Expanding equation (1) at f_o results in Z_g when ℓ_3 is terminated by a capacitor. Rearranging the resulting equation with regard to Z_1 gives us the solution for the following final requirement:

$$Z_1 = \frac{-Z_{o2}Z_{o3}\alpha_3(Z_{o2}\alpha_2 + jZ_g)}{Z_g(Z_{o2} - Z_{o3}\alpha_2\alpha_3) - jZ_{o2}(Z_{o2}\alpha_2 + Z_{o3}\alpha_3)}. \quad (7)$$

It is inferred that as long as α_3 is not zero, which represents the zero length of the ℓ_3 transmission line, Z_1 can be found to obtain a desired value of Z_g .

From these three conditions harmonic impedances are combined so that reduced area is achieved and a rough estimate of the area is found as follows:

$$\text{Area}_{\text{Input Matching}} = (l_1 + l_2) \times (l_3 + W) \approx (L) \times (\lambda_1/4 - l_2 + W). \quad (8)$$

By comparing eqn. (2) and eqn. (8), the occupied area can be reduced by deliberate modifications of the transmission lines at harmonic frequencies. As well, Fig. 2 shows the calculated FOM as per the location of ℓ_3 transmission line, and it is observed that the FOM of the size-reduced design is better than the one without the reduction and that the ℓ_3 should be located nearby the left-side of the circuit. Furthermore, the FOM can be improved by this reduction of the occupied area and a proper control on the impedance ratio.

$$\text{FOM} \cong \frac{|Z_3/Z_1|}{(L) \times (\lambda_1/4 - l_2)}. \quad (9)$$

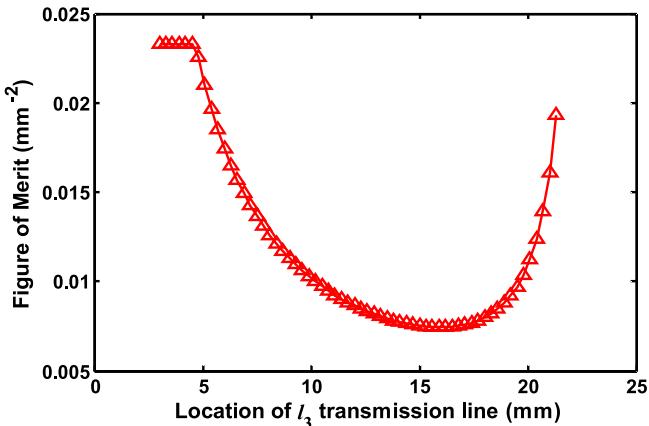


Fig. 2. Figure-Of-Merit (FOM) versus the location of the ℓ_3 transmission line from the left-side port of the input matching network.

3 Implementation of small-sized class-J amplifier

On the basis of the equations presented in the previous section, a class-J amplifier of reduced size is implemented. This amplifier is designed with a commercial GaN CGH40010 transistor from CREE for an output power of 10 W. Mainly intended for the use at 2.1 GHz LTE frequency, the optimum Z_g is selected as $4.2+j2 \Omega$ at 2.1 GHz from the source pull simulation, and Z_R is calculated to be $31-j17 \Omega$ with an ℓ_2 transmission line of 16.8 mm on an RF-35 substrate. Furthermore, ℓ_3 is determined as 3.5 mm at 4.7 mm from the left-side of the network, for an improved FOM.

The output matching network of the amplifier was designed in a similar manner to have the required impedances for the class-J operation.

Specifically, *Condition B* in the previous section needs $Z_L = -jR_L(3\pi/8)$ at $2f_0$; *Condition C* needs $Z_L = R_L(1+j)$ at f_0 ; $Z_L = 0$ at higher harmonics. Alongside these conditions, the optimal R_L is determined as 15 ohm from the multi-frequency loadpull analysis from 2.0 GHz to 2.5 GHz. Then, aforementioned equations are solved with $f_0 = 2.1$ GHz resulting in de-embedded output impedances: $Z_L@f_0 = 12.4 + j15.6$, $Z_L@2f_0 = 0.8 - j12$. Fig. 3 shows the measured output matching result over the desired bandwidth of 2.0 GHz~2.5 GHz after de-embedding of output capacitance, on which loadpull contours of 68% efficiency at 2.0 GHz and 2.5 GHz are overlaid. It is shown that the matching condition satisfies the class-J operation and fairly resides near the high efficiency area over the given bandwidth. Fig. 4 shows the implemented amplifier, where the size of the shaded box represents the amount of shrunk area after the size reduction. Table I summarizes the comparison between the reduction in the area of the designed amplifier and the area of a conventional amplifier without reduction. As a result of the size-reduction, the final area of the amplifier is reduced by 35% as compared with that of a conventional amplifier of similar power [5]. In Fig. 5, the implemented power amplifier shows a peak efficiency of up to 71.3%, an output power of 40 dBm, and more than 55% of efficiency for 300 MHz bandwidth. Thus, it is obvious that this power amplifier performs as well as conventional class-J amplifiers even with a significant amount of size

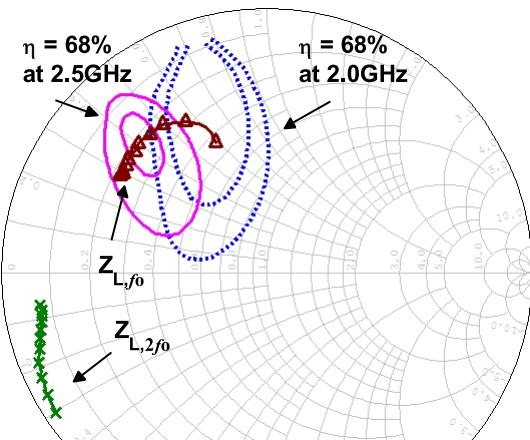


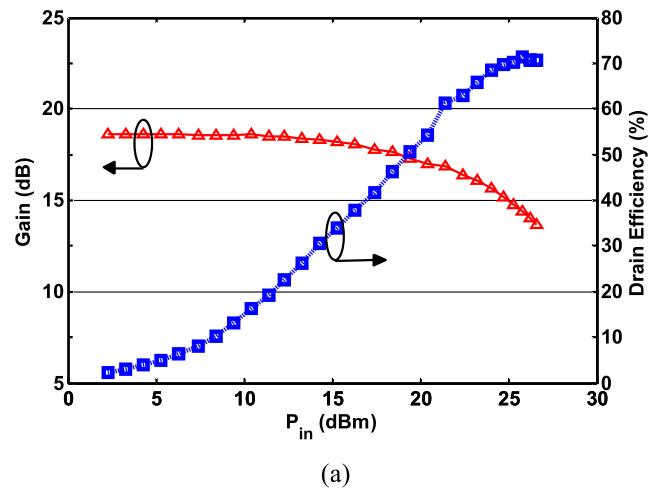
Fig. 3. Implemented class-J output matching network (2.0 GHz~2.5 GHz) overlaid with the 68% efficiency contours from loadpull simulation.



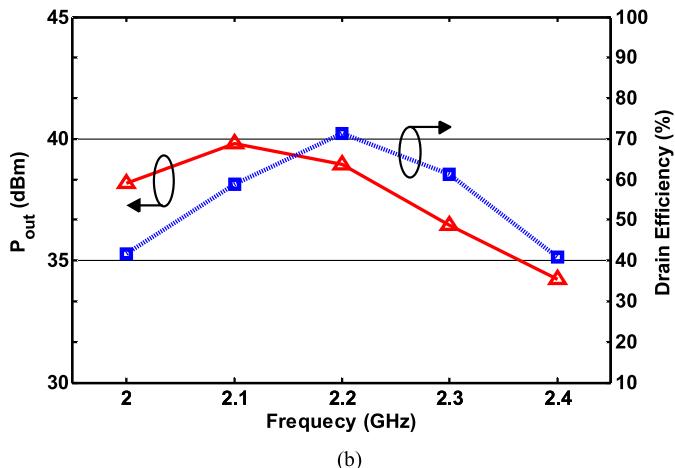
Fig. 4. Designed 10-W class-J amplifier showing the amount of size reduction.

Table I. Reduction in the amplifier's circuit area through the analysis of the harmonic voltage reflections.

	Input Matching (mm ²)	Output Matching (mm ²)	Total area (mm ²)	Size Reduction (%)	Figure of Merit (mm ⁻²)
Conventional Amplifier[6]	28.0 × 42.6	34.5 × 42.6	66.1 × 42.6	-	0.018
This work	28.0 × 27.8	34.5 × 27.8	66.1 × 27.8	35	0.022



(a)



(b)

Fig. 5. Measured performance of the class-J amplifier results (a) Power swept result at 2.1 GHz (b) Frequency response ($P_{in} = 27$ dBm).

reduction.

4 Conclusion

In order to design small-sized power amplifiers for micro-sized radio stations, it is advisable to develop a systematic method rather than simply placing small-sized components. This paper presents an analytic method to combine DC, fundamental, and second harmonic impedances at matching networks of power amplifiers. A class-J power amplifier with this method is designed and evaluated to produce an output power of 10 W at 2.0–2.5 GHz bandwidth. The result of this study shows that this amplifier occupies an area that is 35% less than that occupied by a conventional amplifier. The

amplifier exhibited a maximum efficiency of 71.3%, more than 55% for 300 MHz, and a peak power of 40 dBm. These values are good enough to support the validity of the suggested method.

Acknowledgments

This work was supported by the National Research Foundation of Korea (NRF) grant funded by the Korea government (MEST) (no. 2011-0009106).