A 2.5 W LDMOS MICROWAVE TOTEM-POLE PUSH-PULL RF POWER AMPLIFIER

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RF push-pull power amplifiers (PA) typically use a transformer or balun at the output to combine the two anti-phase signals. The push-pull PA described here uses a totem-pole output stage consisting of two N-channel FETs; a source-follower to supply current to a 50 Ω load, and a common-source amplifier to sink current. The source tab of the source-follower cannot be grounded to dissipate heat in the normal way, so a ceramic spacer transfers the heat to an aluminum heatsink with minimum stray parasitic components.

A prototype PA was designed for 700 MHz. Under continuous wave (CW) excitation, it achieved a 1 dB compression point (P_{1dB}) of 33.9 dBm (2.46 W) at an efficiency of 46.5%. The second harmonic distortion (HD_2) suppression was 43.6 dB relative to the fundamental. These results indicate that the PA was operating in a push-pull mode.

INTRODUCTION

Push-pull RF PAs generally consist of two identical single ended amplifiers operate in anti-phase, with one amplifying the positive signal excursion, and the other the negative excursions. Anti-phase splitting and combining networks based on: wound transformers [1], or baluns [2] [3] are needed at the input and output, which can take up significant board space. The RF Energy Alliance predict that PA volume will shrink by up to 80% in the future [4]. This reduction in size is only possible without transformers and baluns.

At low frequencies both positive (i.e. N-channel FETs and NPN bipolar transistors) and negative (i.e. P-channel FETs and PNP bipolar transistors) polarity device are available. These are often available in complementary pairs making the push pull amplifier shown in Fig. 1 (a) possible, where Q_3 is an NPN transistor and Q_4 a PNP.

At microwave frequencies, performance of negative devices degrades [5] so the transformer coupled push-pull amplifier shown in Fig. 1 (b) is common. This amplifier uses two NPN bipolar transistors (Q_5 and Q_6) or two N-channel FET.

A transformer free alternative is the totem-pole amplifier based on two positive polarity devices (Q_7 and Q_8) as shown in Fig. 1 (c). This has been demonstrated at low RF frequencies with both active input drivers [6] and transformers [7] for the upper FET. A broadband bipolar MMIC

totem-pole amplifier with an active input driver for VHF was described in Reference [8], but resulted in a low efficiency of 14% at 100 MHz with 20 dBm (100 mW) output power (P_{OUT}).

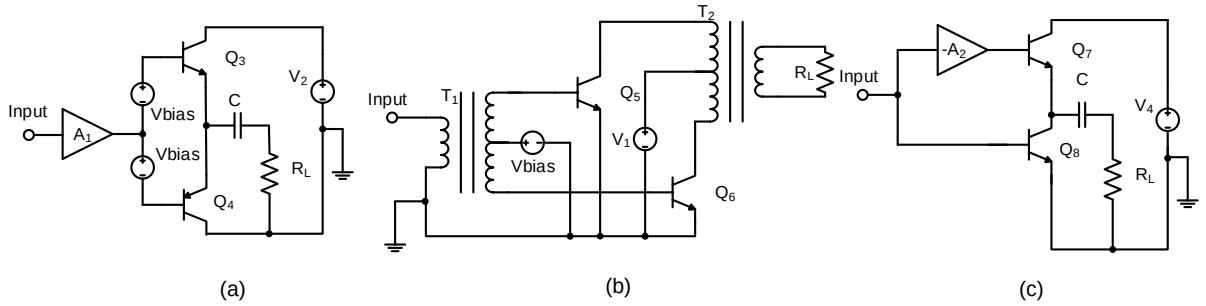


Fig. 1. Push-pull amplifier schematics (a) complementary bipolar, (b) transformer coupled and (c) totem pole

The totem-pole push-pull amplifier in this work is different to that previously described in the literature [6] [7] since it uses a reactive matching network at the input of the upper device. Previous work with GaAs devices [9] has produced a 230 mW P_{OUT} PA. Here, P_{OUT} is increased to 2.46 W with LDMOS devices.

TOTEM-POLE PUSH-PULL AMPLIFIER

The complete schematic for an N-channel based totem-pole push-pull amplifier is shown in Fig. 2. The RF input signal is first split with a Wilkinson power splitter and the resulting output signals gain and phase aligned before applying to the active devices (Q_1 and Q_2), so that they operate in anti-phase. Q_2 is a common source amplifier for sinking current from the 50 Ω load and Q_1 is the source follower to supply current to the load. Variable Resistors R_1 and R_2 set a class B bias.

The low input impedance of Q_2 is matched to 50 Ω with C_2 and L_2. The input impedance of Q_1 is low relative to its source terminal, but high when referenced to ground. Input matching is with C_1 and L_1, which form a voltage step-up network. Source-followers have slightly less than unity voltage gain, but because of their high input impedance, have a large current gain. When the voltage step-up nature of C_1 and L_1 is combined with the high current gain of Q_1, overall power gain is achieved.

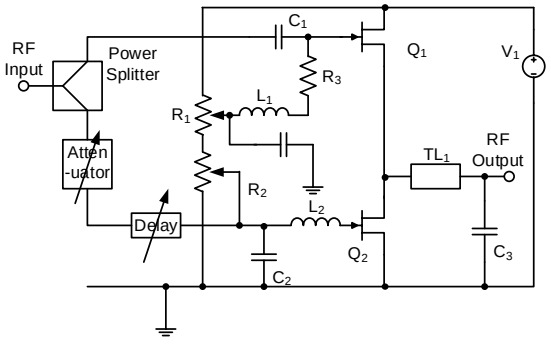


Fig. 2. Totem-pole amplifier schematic

Q_1's input impedance relative to ground (Z_1) is a function of its common-source input impedance (Z_{11}), the load presented to its source (Z_L) by the matching network formed by transmission line TL_1 and C_3, and its transconductance (g_m):

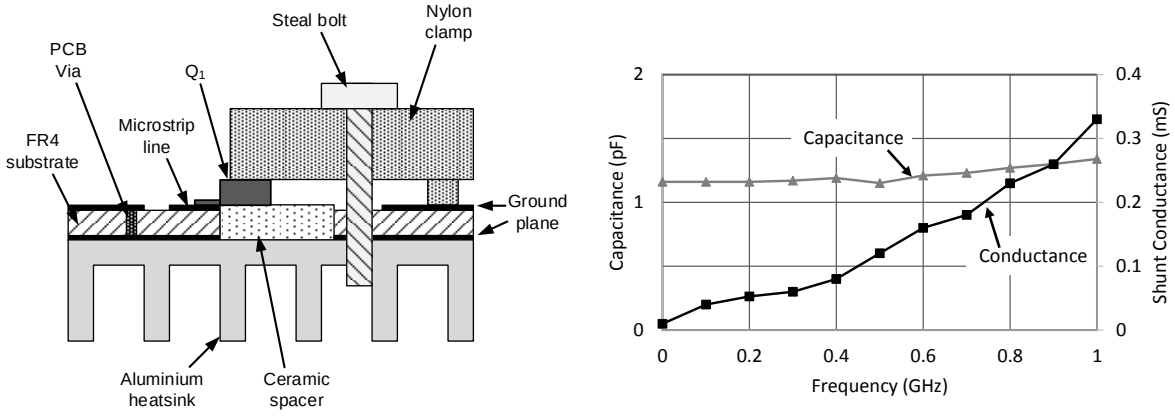
$$Z_1 = Z_{11} \cdot (2 \cdot Z_L \cdot g_m + 1) \quad (1)$$

It is assumed that Q_1 and Q_2 are the same device, although this does not necessarily have to be true. Resistor R_3 improves stability since feedback can occur through Q_1's gate-source capacitance (C_{gs}).

Since the two halves of the amplifier are asymmetric, their phase shifts will be different. The Delay element in Fig. 2 is adjusted for correct push-pull operation. To suppress the HD_2, the gain of the two halves must also be matched, a variable attenuator performs this task.

SOURCE-FOLLOWER COOLING

High power RF transistors are generally bolted or soldered to a heatsink via their source terminal to dissipate excess heat. In a totem-pole PA, Q_1's source carries an RF signal, so cannot be. It must however still be thermally connected to the heatsink, but not electrically. The amplifier described in Ref. [10] used alumina insulators between the active devices and a heatsink. Here, a MPC101020T ceramic insulator from Amec Thermasol transfers the heat dissipated by an LDMOS PD85004 RF power transistor to the heatsink. To ensure minimum stray capacitance a nylon block clamps Q1 to the ceramic insulator and in-turn to the aluminium heatsink as shown in Fig. 3 (a). Thermal paste is used between all contacting surfaces.



(a) (b)

Fig. 3. (a) Mechanical arrangement of heatsink, (b) capacitance and shunt conductance of ceramic spacer

The junction temperature (T_j) of Q_1 must not exceed 150°C. Assuming an ambient room temperature (T_{amb}) of 25 °C, the maximum temperature rise (ΔT) is 125 °C [11]:

$$\Delta T = T_j - T_{amb} \Delta T = T_j - T_{amb} \quad (2)$$

The maximum power dissipation (P_{diss}) for Q_1 can be calculated:

$$P_{diss} = \frac{\Delta T}{\theta_{total}} \tag{3}$$

Where θ_{total} is the total thermal resistance from Q_1's junction to free space:

$$\theta_{total} = \theta_{jc} + \theta_{cs} + \theta_{sa} \tag{4}$$

The ceramic insulator had a thermal resistance of 10.2 °C/W (θ_{cs}), and the heatsink 3 °C/W to ambient (θ_{sa}). These are both less than the 21 °C/W junction-to-case thermal resistance (θ_{jc}) of the PD85004. Therefore θ_{total} is 34.2 °C/W. θ_{total} for Q_2 will be different since it dissipates its heat through PCB vias, which can also be calculated [12].

A stray capacitor is formed by the source terminal of Q1, ceramic insulator and heatsink. In reference [10] this stray capacitance was 8 pF. The capacitance here was measured as 1.2 pF by attaching pieces of copper tape to each side of the ceramic spacer. One side completely covered the spacer to represent the ground plane, and the other was approximately the same size as the source tab. By measuring the S_{11} with a vector network analyser, the capacitance and conductance was calculated. The capacitance is reasonably flat as shown in Fig. 3 (b), while the conductance at 700 MHz was 0.2 mS. Both are low enough to have a negligible effect.

AMPLIFIER DESIGN

The amplifier was constructed on a single sided FR4 substrate with ground plane as shown in Fig. 4. A hole was cut to accommodate the ceramic insulator, as indicated in Fig. 3 (a). A resistive attenuator pad was included in one branch of Wilkinson splitter for gain alignment. Phase alignment was with calibrated lengths of cable and SMA adapters.

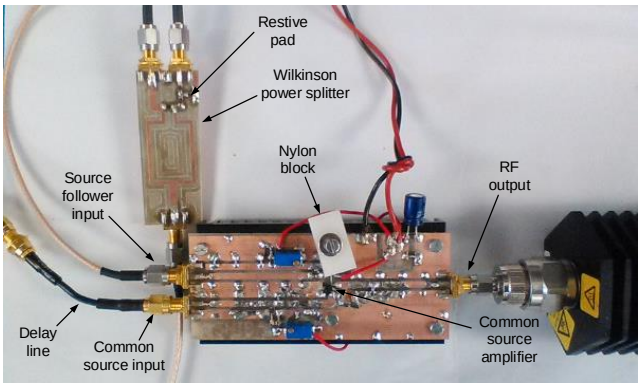


Fig. 4. Photograph of amplifier

Using the datasheet as a guide, the input and output matching networks were calculated. An optimum Z_L of 6.3 + 1.7j Ω was found and Z_{11} 17.4 – 29.6j Ω at the operating frequency of 700 MHz. The g_m from the datasheet was 0.75, resulting in a Z_1 of 257 – 267j Ω. R_3 was chosen as 360 Ω, reducing the source-follower input impedance to 183 – 77j Ω.

PRACTICAL RESULTS

With a CW signal applied to the PA, the gain and phase imbalance between the two paths were set to a given value and input power swept. P_{OUT}, efficiency and HD_2 suppression were recorded. The P_{1dB} sweep is shown in Fig. 5 (a). The efficiency and HD_2 suppressions had similar profiles as shown in Fig. 5 (b) and Fig. 5 (c) respectively.

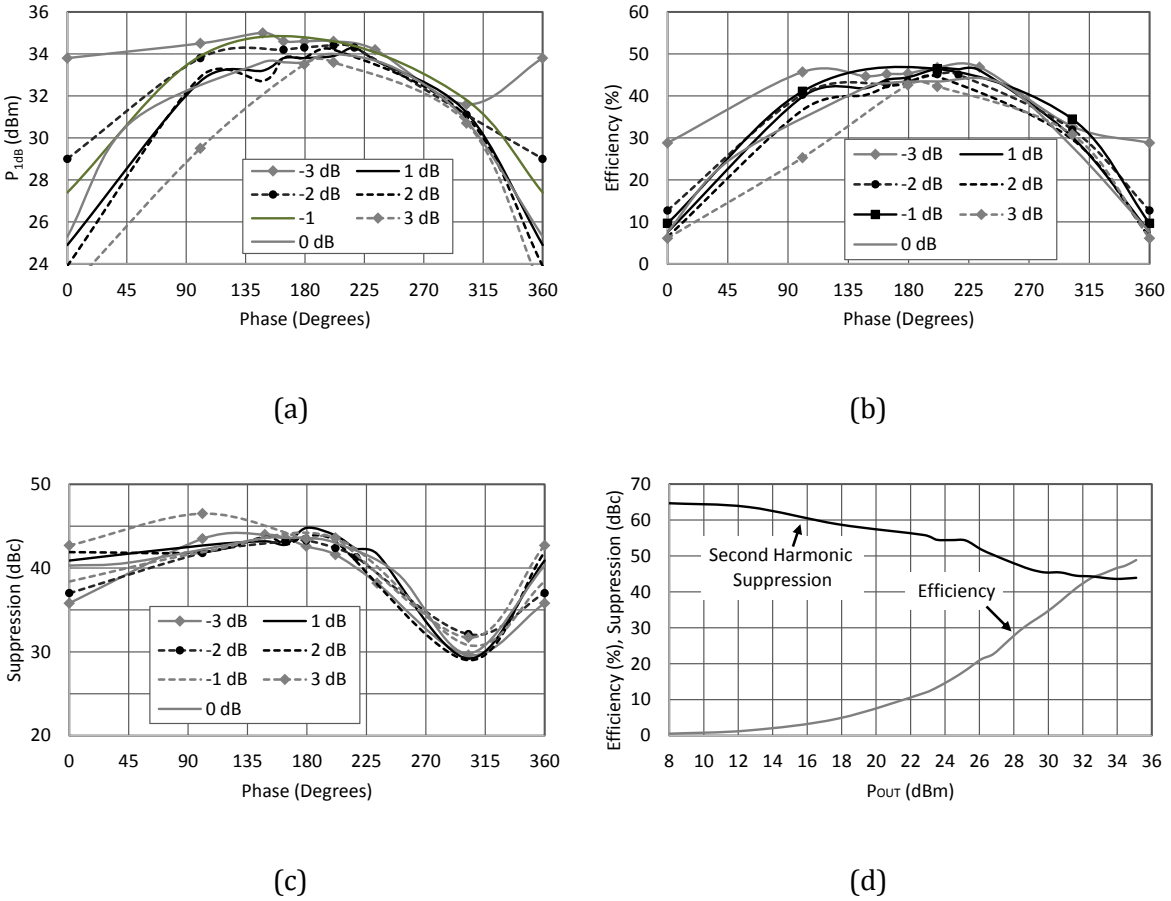


Fig. 5. (a) P_{1dB}, (b) efficiency and (c) HD_2 at the P_{1dB} verse gain and phase imbalance. (d) Power sweep at optimum gain and phase setting

In Fig. 5 (a), (b) and (c), a negative gain indicates attenuation in the common-source path and a positive the source-follower. The optimum gain and phase imbalance were determined to be -1 dB and 148° respectively, resulting in a P_{1dB} of 33.9 dBm (2.46 W) and efficiency of 46.5%. Therefore the total PA power dissipation was 2.82W, resulting in a Q_1 P_{diss} of 1.42 W. Using (2) and (3), ΔT was 48.4°C and T_j 73.4°C, well below the 150°C maximum. The HD_2 suppression was 43.6 dBc relative to the fundamental. A sweep of efficiency and HD_2 suppression against P_{OUT} is shown in Fig. 5 (d). There is very little comparable research, but it is compared to closest published work in Table 1.

Table 1: Comparison of totem-pole amplifiers with CW signal

Reference	Measurement Frequency (MHz)	P_{OUT} (W)	Efficiency (%)
[5]	8000	0.01	20
[6]	1	3.7	58.3
[7]	1-30	560*	77.0*
[8]	100	0.1	14.0
[9]	680	0.23	52.0
This Work	700	2.46	46.5

*estimated from modulated results

CONCLUSION

A totem-pole RF PA is described composed of common-source and source-follower amplifiers. A ceramic insulator transfers dissipated heat from the source-follower to a heatsink. When correctly aligned 46.5% efficiency was achieve at 33.9 dBm P_{OUT}. Under these conditions the HD_2 suppression was 43.6 dBc relative to the fundamental output signal. The high efficiency and HD_2 suppression indicate that this amplifier is operating in push-pull.

ACKNOWLEDGMENTS

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