The Output Impedance of HF Power Amplifiers: Not as expected

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Modern transceivers for the shortwave range offer broadband transmit amplifiers with LDMOS transistors in push-pull circuit and a transformer on the output side for "matching" to a 50 Ω load impedance. In amateur radio circles and even in the catalogue of questions for the license test, it is assumed that the transmitter output impedance, i.e. the internal resistance of the amplifier at the output, is therefore 50 Ω. But is that really the case?

The Thevenin and Norton equivalent sources

The idealization of an RF amplifier as a voltage source with internal resistance or as a current source with internal conductance goes back to the simplification of circuit calculations by "Thevenin" and "Norton" equivalents in the theory of "linear electrical networks". At the terminals of these equivalent sources, the same voltage and current flows at any given load as at the terminals of the circuit that is being replaced. However, the equivalent sources do not describe the real conditions within the replaced circuit.

A well-known application is the Thevenin equivalent voltage source of a battery, Figure 1: The internal resistance is the quotient of open-circuit voltage and current due to a short circuit at the terminals, which should not be caused on a battery. It is better to measure the open-circuit voltage U_0 without load resistance, while with any load resistance a voltage drop at the load resistor is measured, from which the internal resistance can be calculated as $R_I = \left(\frac{U_0}{U_L} - 1\right) R_L$. If the load resistor is equal to the internal resistance, half the short-circuit current flows (at $R_L = 0$) and the voltage at the load resistor U_L is half the open-circuit voltage. In this case, the equivalent voltage source delivers the maximum power that can be discharged (or is available): $P = U_L^2/R_L = U_0^2/(4\ R_I)$. The efficiency of the equivalent voltage source in this case is only 50% because the terminal current also flows through the internal resistance, which is the same as the load resistance. Batteries are therefore usually operated with a much higher load resistance, so that less energy stored in the battery is lost as heat, i.e. the efficiency is higher.

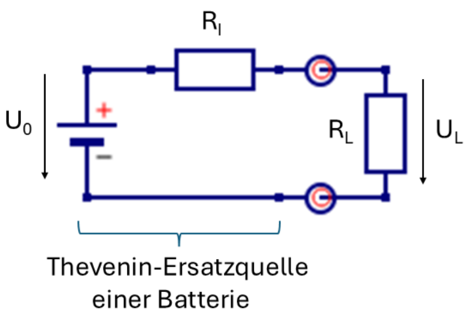

Figure 1 Thevenin equivalent voltage source of a battery with load resistor

In principle, the concept of the substitute source can also be applied to linear AC networks, e.g. passive circuits, such as filters or attenuators, but also active circuits, such as amplifiers, as long as limiting and saturation effects do not set in at high levels. In contrast to DC circuits, high-frequency circuits add reactive components to the dissipative resistances in the load and the equivalent sources, so that, for example, a Thevenin equivalent voltage source delivers the maximum power if the load impedance also compensates for the reactance of the internal resistance at the same time – this is then the "conjugated complex impedance matching".

The simplest case of an RF current source that can be represented as an equivalent voltage source is the laboratory signal generator, which has a switchable attenuator on the output side. The open-circuit voltage of the equivalent source depends on the setting of the attenuator, but the internal resistance is always close to 50 Ω. On a transmission line, the signal generator therefore generates the same amplitude of the forward travelling wave at any load and a reflected wave due to mismatch of the load is completely absorbed by the internal resistance.

In the case of the laboratory signal generator, there is effectively a 50 Ω physical resistance between the actual signal source and the output terminals in the form of the attenuator. However, this cannot be afforded with a power amplifier, as a large part of the generated RF power would be converted into heat in this resistor. RF power amplifiers with output powers above a few watts are usually built with transformation circuits on the output that are intended to provide maximum output power at a 50 Ω load. Therefore, it is often assumed that such a power amplifier can be described by an equivalent voltage source with 50 Ω internal resistance.

Unfortunately, this is not so easy – the way the amplifier transistors work plays a decisive role here. This will be explained with an example: Figure 2 shows a simple amplifier circuit operating in class A, with a FET as the active element; Figure 3a shows the appropriate idealized output I-V characteristic of the transistor. In its operating point, the FET is idealized as a Norton equivalent source for alternating current, where the current source on the drain side is controlled by the gate voltage (proportionality factor *s*) and there is a resistor parallel to the current source. In the amplifier circuit, the FET is controlled by the voltage U_{GS} at the gate and loaded by a resistor R_L on the drain side. With a DC voltage at the gate $U_{GS=}$ = 0 V, a bias operating point AP is set, which marks the point of the corresponding output I-V characteristic curve for the applied drain DC voltage of 12 V. The slope of this output characteristic curve corresponds to the conductance G_{DS} of the drain resistance in the Norton equivalent source of the FET. The AC voltage $U_{GS\approx}$ at the gate shifts the I-V output characteristic curve up and down. The operating point runs along the red operating line, the slope of which corresponds to the negative conductivity of the connected load. This results in an AC voltage $U_{DS\approx}$, which is also applied to the load resistor if the reactance of the decoupling capacitor is negligible. The resulting alternating current $I_{D\approx}$ also flows only through the load because of the choke, but here with the opposite sign or direction.

The picture shows the maximum gate voltage – with stronger excitation the transistor would be driven into limitation and the sinusoidal voltage and current curves would be distorted: The transmission characteristics of the amplifier would become non-linear. The AC power delivered to the load is thus the maximum possible in linear operation at the control of 2 V_{pp} at the gate given here with $P = U_{DS\approx} \cdot I_{D\approx} = 0.5$ W. It can be seen that the selected slope of the operating line is the same as the slope of the output characteristic curve, only inverse: This is exactly the case of the "impedance" matching of the load to the internal resistance of the equivalent source $R_L = 1/G_{DS}$. Therefore, the power achieved is also the available power, which can only be delivered in matched condition. In addition, the voltage gain also becomes maximum with this adjustment: $V_U = 20$ V/2 V = 10 or 20 dB. Operating point, gain limit and impedance match are typical for a small signal amplifier that can be correctly described by an equivalent source. The equivalent source is the controlled current source in Figure 2b together with the parallel conductance G_{DS}; the equivalent Thevenin source has the open-circuit voltage $s \cdot U_{GS\approx} / G_{DS}$ and in series connection the internal resistance $R_I = 1/G_{DS}$. In practice, therefore, it is said that the real circuit has an internal resistance. In small-signal circuits, this internal resistance on the output side is usually determined by a reflection factor measurement (S22) at the output of the amplifier.

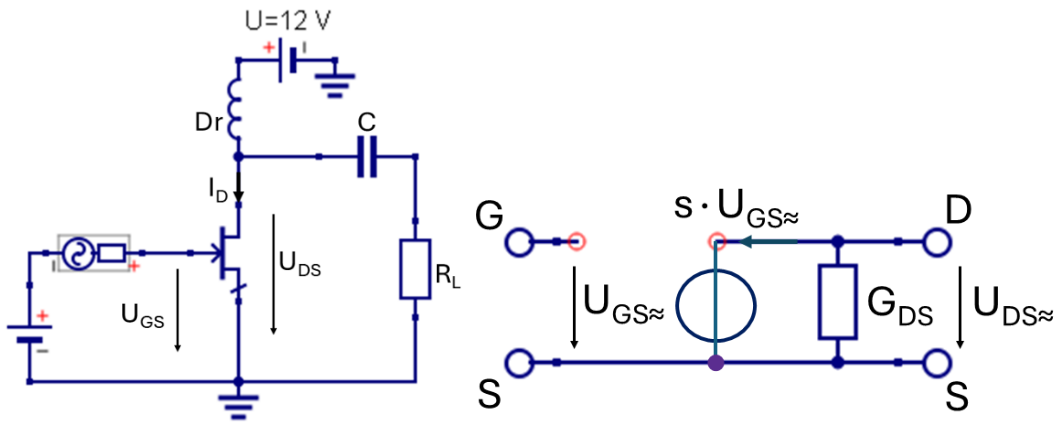

Figure 2 (a) Amplifier circuit with FET, (b) AC voltage equivalent circuit diagram of a FET with controlled Norton equivalent source; ≈ means AC

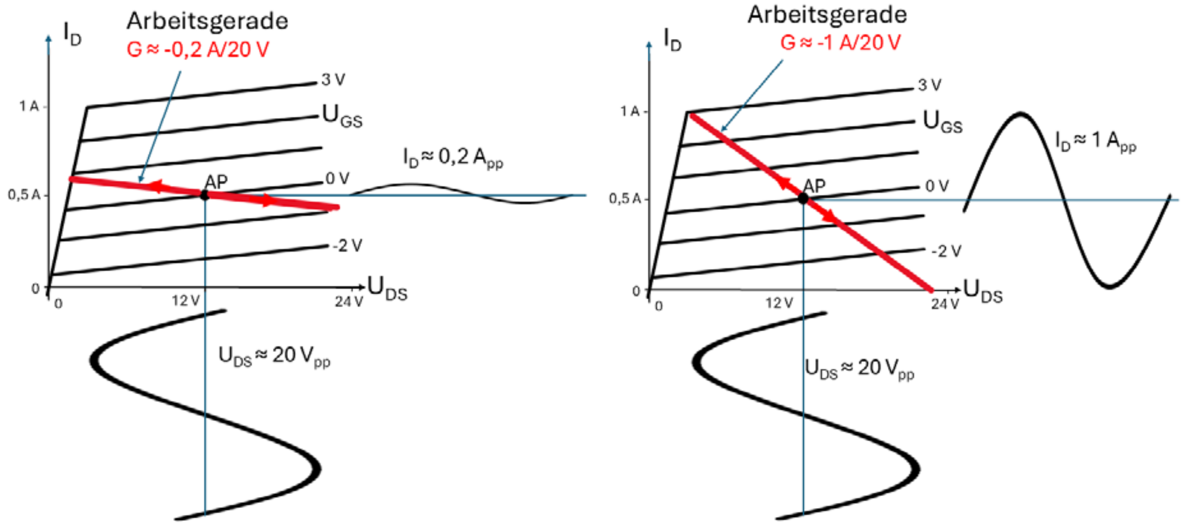

Figure 3 Idealized output I-V characteristic of a FET with operating lines in class A for (a) small signal operation with about +/- 1 V at the gate, (b) large signal operation with about +/-3 V at the gate.

Large-Signal Behavior

Because of the limitation to <u>relatively</u> small voltage / current excursions in the I-V characteristic, we call it small-signal operation, where the rules of the "linear networks" apply. The problem with this is that the circuit with an impedance-matched load does not take advantage of the transistor's capabilities: With a lower load resistance value, the amplifier can deliver considerably more AC power, although not impedance-matched. The principle can be seen in the initial I-V characteristic Figure 3b. The load resistance with a higher conductivity is represented by a steeper operating line. This allows the transistor to be excited over a much larger voltage range at the gate without overloading or limiting effects. At the same voltage swing at the drain as in the small signal case, a much higher drain current can flow and the product gives a much higher output power; in the example, the load resistance is 5 times smaller than with match condition, the drain current is 5 times larger and thus the power delivered is also 5 times greater. However, the voltage gain is also 10 dB smaller: The amplifier still works in linear mode in large-signal operation. It is also shown that for the $U_{GS\approx}$ control given here, the maximum achievable RF power is delivered with the operating line entered in the plot (corresponding to a conductivity of the load resistor of 1/20 S); with even greater excitation at the gate, the optimal operating line would be even steeper, and the optimal conductance value of the load resistor would be even greater.

Maximum achievable output power is actually a characteristic that stands for power matching of an equivalent source, where the internal resistance corresponds exactly to the load resistance. So, have we also found the internal resistance of the equivalent source with the optimal load resistance? Is there an argument against the fact that the conductivity of the optimal load is the greater the power level is, and the internal resistance would therefore not be a constant?

What does the practice look like?

We have to test this practically with a real power amplifier. The PA of my FT450 offered itself as a good-natured test object, which remains stable with any passive load of input and output, i.e. does not tend to self-excitation and oscillation. So, the internal input signal was unplugged at the first stage of the amplifier and the output line of the PA was disconnected directly behind the push-pull transformer on the PCB in order to connect a coaxial cable for measurement. In Fig.4 you can see the power amplifier of the PA (a transistor and a part of the transformer at the bottom right) with the RG174 cable on the disconnected strip line. Above this point is the filter group, directional coupler, tuner and transmit-receive switching relay up to the PL output socket. Because of the strong frequency response of this path, it is not useful to measure the output impedance of the PA directly at the output socket.

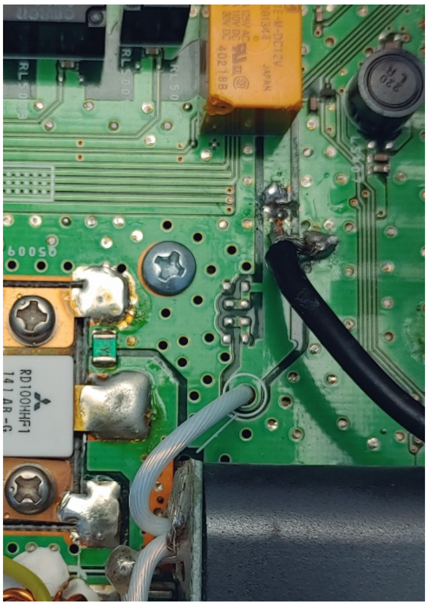

Figure 4: Decoupling at the output of the PA of my FT450

Measurement of the initial reflection coefficient

As a first measurement, the reflection coefficient S22 was measured at the PA output with the help of a NanoVNA, while the driver stages were shut down (without input signal). The VNA "sees" the impedance on the output side, which in our example circuit would correspond to the conductance G_{DS} of the equivalent current source.

Without applied DC operating voltage, the measured curve of the reflection coefficient comes close to a lossless parallel resonant circuit, Figure 5. The capacitance in this resonant circuit comes mainly from the drain-source capacitance of the LDMOS transistors and the inductance from the push-pull transformer. When the operating DC voltage is applied, the measurement curve shifts inwards, which corresponds to a resistor connected in parallel. At about 3.4 MHz, the resonant circuit is in resonance, and we can read the active resistance – here about 290 Ω.

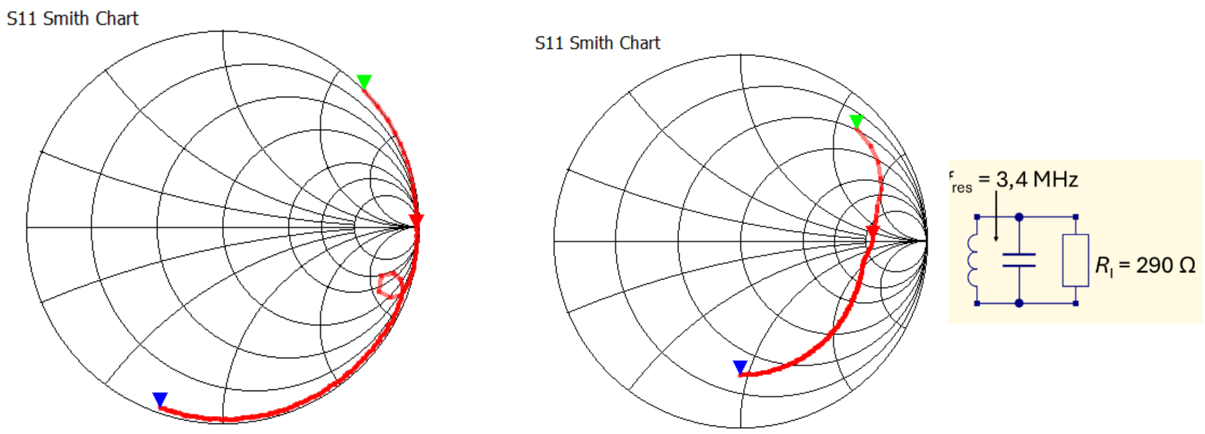

Figure 5 Reflection coefficient from 1 MHz to 30 MHz at the PA output, (a) without operating DC voltage, (b) with operating DC voltage. Red marker at 3.4 MHz.

However, the measurement with -10 dB_m as a test signal from the NanoVNA only excites the power amplifier transistors in the mV range, which would be perfectly fine for a small signal amplifier. With the help of a directional coupler and an amplifier, the output-side reflection coefficient S22 was measured at a higher level /1/, but only up to about 1-2 W of power running towards the PA output. At higher power, the measurement gives an incorrect picture, because the linear range of the current and voltage excitation is left due to limiting effects in the output I-V characteristic.

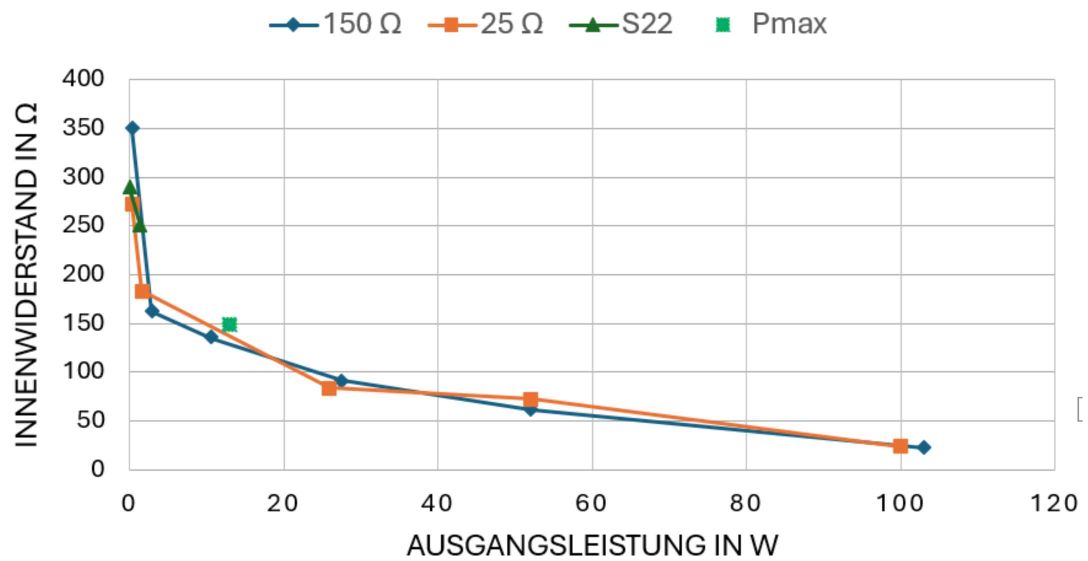

Figure 6: Measurement results for the internal resistance of the PA plotted over the nominal output power of the PA at 50 Ω.

Tests with high output power of the PA

Similar to a battery, an internal resistance of the PA can be inferred by measuring the output voltage at different load resistors. The load cases were chosen so that the PA still works almost linearly up to high levels, so the load resistances should remain close to 50 Ω. In order not to endanger the PA by exceeding limit values for the transistors and to be able to go up to a full 100 W output power, VSWR = 3 should not be exceeded. Accordingly, thick-film resistors on heat sinks with 150 Ω, 25 Ω and 50 Ω were used as load resistors. A NanoVNA was used to excite the PA with post-amplification of the output signal from port 1 (S11) in order to achieve the necessary input power to the PA up to full level. Part of the preamplifier was a bandpass filter to suppress the harmonics of the VNA's output signal. The voltage U_L at the load resistor R_L was first measured directly with an oscilloscope. It was found that the voltage curve still showed strong deviations from an approximate sine wave shape (i.e. high harmonic component), depending on the load resistor used, so that voltage comparisons for determining an internal resistance seemed too imprecise. The fundamental wave contained in the voltage was instead measured with the VNA itself by feeding the signal into port 2 (S21) or alternatively with a spectrum analyzer (tiny SA). A directional coupler was used as a measuring coupler: At its port **a**, the forward travelling wave leading up to the load is coupled out with high coupling loss; for control, the reflected wave can be observed at port **b**. The entire circuit arrangement is shown in Figure 7.

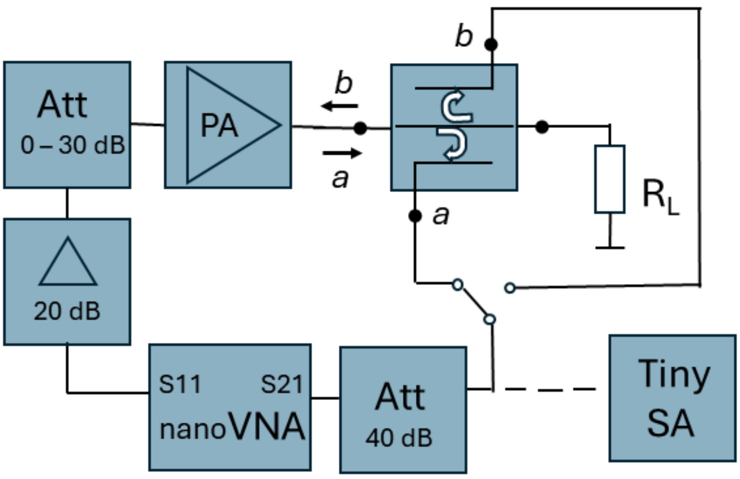

Figure 7 Measurement setup for determining an internal resistance of the PA at high output power

The measurements were carried out at 3.5 MHz, where, according to the measurement of the reflection coefficient, the internal resistance of the PA can be assumed to be almost pure dissipative (real). This assumption has also been confirmed for high output powers; with pure resistive load impedance, the calculations become easier.

The ratio of the measured voltages U_L at different load resistances allows the internal resistance of the PA to be determined, similar to the battery by assuming an equivalent voltage source as in Figure 1:

With the voltage divider rule, the voltage at equivalent voltage source terminals is

$\frac{U_L}{U_0} = \frac{R_L}{R_L+R_I}$ with the sought-after internal resistance R_I of the source.

If the PA is operated on a transmission line, the voltage U_L can be split into the voltages of two waves travelling forward (outgoing) and backward (reflected). The voltage of the outgoing wave U_h results from the total voltage U_L and the reflection coefficient *r* to

$$\frac{U_h}{U_0} = \frac{U_L}{U_0} \cdot \frac{1}{1+r} = \frac{R_L}{R_L+R_I} \cdot \frac{1}{1+r} .$$

Given the reflection coefficient of the load, both the open-circuit voltage U_0 and the internal resistance R_I are unknown in the equation. We therefore need a second equation with other load resistance to solve for the internal resistance. This leads to the quotient ***c*** of two voltages U_{h1} and U_{h2} from the measurement with the reflection coefficients r_1 and r_2 of the load but constant excitation of the driver stages of the PA. Since both voltages are mapped as a coupling signal "**a**" via the directional coupler, the quotient is identical to the ratio a_1/a_2 of the two measured couplings. Solving the equation for ***c*** for to the sought-after internal resistance R_I for the three load resistors used and two combinations of them results in:

$R_I / Z_0 = (3\,c - 2) / (2 - c)$, für a_1@ $R_L = 150\ \Omega$ und a_2@ $R_L = 50\ \Omega$,

$R_I / Z_0 = (2/3\,c - 1) / (1 - 4/3\,c)$, für a_1@ $R_L = 25\ \Omega$ und a_2@ $R_L = 50\ \Omega$,

with Z0 = characteristic impedance of the transmission line of 50 Ω.

In the measurements, the input level at the PA was varied in such a way that output power was generated at 50 Ω load from less than 0.2 W to 100 W. Two internal resistances were calculated for each power level. The results are shown in Figure 6. In addition to these results, the load resistors for 1.3 W and 13 W are also entered, for which a maximum power output P_{max} was determined experimentally and which can also be interpreted as internal resistances; Figure 8 shows the measurement results for one of the two data points.

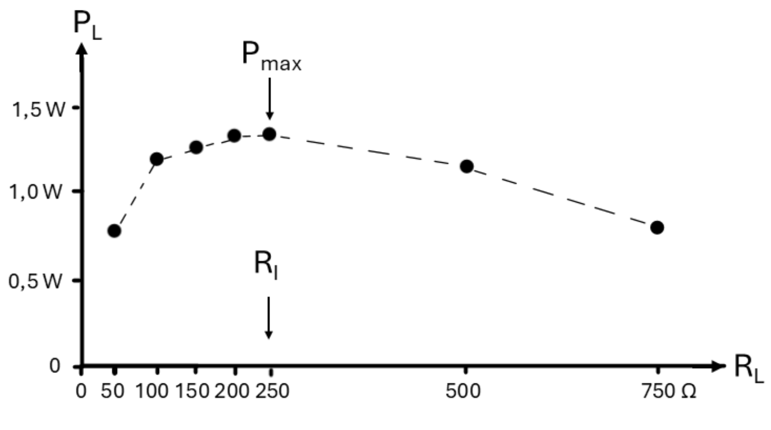

Figure 8: Measured power output of the PA depending on the size of the load resistor between 50 and 750 Ω based on an output power of 0.7 W at 50 Ω.

Learnings from the results?

The first finding from the data in Figure 6 is that the internal resistance of about 290 Ω obtained from the S22 reflection coefficient of the PA does not contradict the measurement results with variation of the load resistances – at very low powers, the results are recognizably similar.

From the results of the measurements with higher output powers, two astonishing findings can be seen:

Contrary to expectations, the equivalent source model with an internal resistance also seems to be realistic in the area of the large signal control of the PA, since the test provides a satisfactory fit between internal resistances from the measurement with low load resistance and those from the measurement with high load resistance within the framework of the load mismatches with VSWR up to 3. Apparently, internal resistance and optimal load resistance are indeed closely related; in any case, the load resistors determined for maximizing the output power at 1.3 W and 13 W also approximately confirm the determined internal resistances. Crucially, the PA for all output power and all load resistors is largely linear in the tests; in this respect, the determined internal resistance of the PA is a "small-signal" quantity despite the large-signal operation of the PA. This means that the internal resistance loses its importance as soon as current and voltage wave forms of the PA show clear limiting effects.

Secondly, the determined internal resistances depend strongly on the excitation level and thus on the output power; the higher the power, the lower the internal resistance drops. Figure 6 shows that impedance matching at 50 Ω can be expected at about 70 W output power, while at 100 W the internal resistance is significantly lower than 50 Ω. In accordance with the determined internal resistances, the higher output power was measured in all measurements up to 50 W with the 150 Ω load compared to the 50 Ω termination and at 100 W with the 25 Ω load. The low internal resistance at 100 W is surprising, as one would rather assume that the PA output matching circuit would be tuned to 50 Ω at maximum power. This tuning in particular can be different for each amplifier, while the drop in internal resistance with increasing output power is in principle related to the operating line in the output I-V characteristic of the power transistors that corresponds to the level of excitation.

We have now determined an internal resistance of the PA, but the meaning is limited because the validity of our equivalent source model is limited to a linear operating state: The second element of an equivalent source, open-circuit voltage and short-circuit current, is completely missing, because at higher output powers both are outside the linear operating state of the PA. Even the output internal resistance only applies to a limited range of the load impedance around 50 Ω. However, if our model of the equivalent source is not completely valid, the theoretical statements about the properties of an ideal equivalent source lose validity, in particular the repeatedly cited limitation of the efficiency to 50% at maximum power output. This may explain the apparent contradiction that power amplifiers (not the entire PA or transceiver) with modern LDMOS transistors achieve over 70% total efficiency (RF power over sum of all input powers) at full gain and power match, e.g. /2/.

There remains the interesting question of what happens within the PA in the event of mismatch, if there is no internal physical resistance to absorb the returning power. To answer this, you have to look at the coarse of current and voltage at the power amplifier transistors; this is particularly exciting, since modern transceivers react to mismatch by power control in the PA and the combination with an automatic matching network (tuner) creates even different conditions.

References

/1/ K. Solbach, "SWR meter becomes directional coupler"

/2/ https://www.ampleon.com/products/general-purpose-wideband/12.5-14-v/BLP5LA55SG.html