RF Amplification with Triodes [1] *

Dr. Rudolf Cantz

1953

1 Introduction. Cathode current negative feedback with tapped input circuit

Triodes are now often used for the amplification of very high frequencies. The grounded grid circuit has been used most often in single stage preamps, where you can do without neutralization. In addition, one also finds the grounded cathode circuit which must be neutralized in general very accurately, if it is to operate stably. Recently, a special combination of a grounded cathode followed by a grounded grid stage has been widely used: the Cascode circuit, which is implemented with a double triode. Here, the output circuit of the first stage is so heavily damped by the low impedance input of the following grounded grid stage that the first stage is easy to neutralize.

The most important variables to consider in the RF-amplification of short waves are the noise figure, gain and bandwidth. From fundamental considerations, it follows that the optimum noise figure and the corresponding optimal antenna coupling to the grounded cathode circuit and grounded grid circuit are the same [2]. This ”noise-optimized” realization is different in both cases from the ”power gain optimized” realization which provides the highest gain. In the grounded cathode circuit, the antenna is known to be more coupled for ”Optimum Noise” (**ON**) than for ”Optimum Power gain” (**OP**) [2]. As already mentioned, **ON** antenna coupling is equally applicable to the Grounded Grid (**GG**) circuit and Grounded Cathode (**GC**) circuit. However, for higher **OP**, the antenna coupling must be stronger because of strong damping by the tube. This wide difference between **ON** and **OP** represents a further disadvantage of **GG** circuit in addition to the well-known disadvantage of a low input resistance. If such a stage is set for lowest noise figure, power gain will be significantly below optimum. Also, if you use an antenna cable, this **ON** optimization produces a mismatch to the cable's characteristic impedance. This results in reflections at the preamp which, if also made at the antenna can lead to some reception inaccuracies, for example, with the appearance of multiple edges in a television picture.

Receiver development for FM-band broadcasting in the 100 MHz band, gave occasion to examine the various options in more detail about how a low-noise RF stage can be built with a single triode. After initial experiments with a **GG** circuit, the author thought of returning the anode circuit neither to the grid nor the cathode, but to a grounded tap at the grid-cathode tank circuit. While pictures. While pictures 1.1 and 1.2 represent the concepts of the well-known **GC** and **GG** circuits, Figure 1.3 shows the concept of the new circuit.[1] It is seen that the Cathode current flows through the lower part of the resonant circuit, thereby producing a kind of current feedback which is, however, then correspondingly lower than in the effective **GG** circuit current feedback. The circuit shown in Figure 1.3 therefore has properties that, depending on the choice of the tapping point, are somewhere between those of **GC** circuit and those of the **GG** circuit.

It is possible to set the tap so that **ON** and **OP** coincide, or are at least so close together that the lowest possible noise optimum occurs near the lowest gain loss and the receiving antenna cable has a reflection-free termination at the receiver.

---

* Original German language article scans, review and translation proofing by Hans Knoll and Dietmar Rudolph. Translation by Joe Sousa with help from Microsoft and Google translators. February 4th 2012

[1] The term “Zwischenbasis-Schaltung” — “Intermediate Basis Topology” is not yet used in this paper. However it appears in a 1953 TELFUNKEN paper (“Röhren- und Halbleitermitteilungen” No. 550 502) written by R. Cantz: “Die Schaltung der Vorstufe im UKW-Rundfunk-Empfänger”.

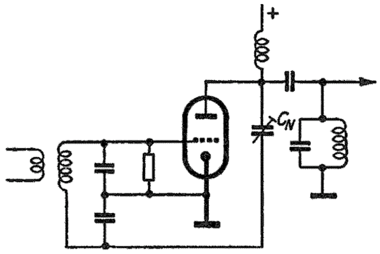

Bild 1. Hf-Stufe in Katodenbasisschaltung, neutralisiert

Figure 1.1: Neutralized RF-stage grounded-cathode (**GC**)

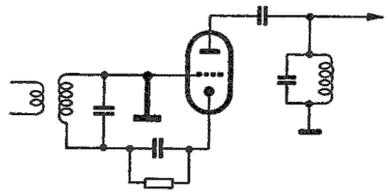

Bild 2. Hf-Stufe in Gitterbasisschaltung

Figure 1.2: RF stage grounded-grid (**GG**) circuit

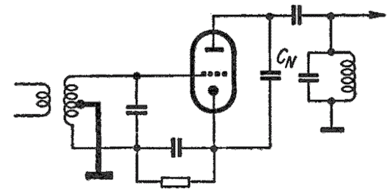

Bild 3. Hf-Stufe in Gegenkopplungsschaltung, neutralisiert

Figure 1.3: Neutralized RF stage in negative feedback topology (i.e. Intermediate basis RF circuit)

The value of input impedance of this negative feedback circuit is between the impedance of **GC** circuit and the impedance of the **GG** circuit. In FM-band[2] radio receiver design the designer wants generally only two-gang tuning for the oscillator circuit and the input circuit of the mixer. The input stage will then stay fixed-tuned in the center of the FM-band radio band and must be coupled with the necessary antenna bandwidth of 10 to 15 MHz. This would require an additional damping resistor to the grid circuit of the **GC** preamp to get this bandwidth, while on the other hand, the **GG** circuit delivers a usually a much too large input bandwidth. While the former would result in an increase in figure, it would, in the latter case, worsen the image rejection. The new negative-feedback circuit allows the implementation of the required bandwidth, without degrading the noise figure.

Of great importance for the Hf-stage is the immunity from self-oscillation, under-damping, and from a shift of the resonance frequency due to anode reaction. Therefore overall stability of the RF stage is increased. The **GG** circuit initially has the advantage that its stability is almost always sufficient, even without neutralization measures because it's input impedance is extremely low. In contrast, the **GC** circuit must be neutralized not only in every case, but the accuracy of neutralization at high frequencies is very stringent, if the stage is designed for greater RF amplification. The circuit shown in Figure 1.3 has a stability and the neutralization difficulty that lie between the **GC** and **GG** circuit. A small neutralization capacity lays between anode and cathode in the new circuit. It's size is less critical than in the **GC** circuit. For a specific tapping ratio it is possible to achieve neutralization with just the C_{ga} and C_{ka} capacitances which lie inside the tube, in addition to wiring capacitance, without having to require an additional external capacitance.

Favorable properties are thus expected from the new negative-feedback. Therefore, the author carried out both calculations and measurements on the new circuits in the laboratory. The results of these analysis are presented below.

2 The input admittance

The conductance between the grid and cathode plays an important role in all further considerations. We proceed from Figure 2.1 for its calculation. If we assume that the input circuit has an input current I_e, and which flows through k and g, and produces a voltage U_e between these points, the input admittance is defined as $G_e = \frac{I_e}{U_e}$. This value includes on the one hand, the effective conductance $G_{\mathrm{kr}} + G_{\mathrm{el}}$ of the tuning circuit, including the electronic input resistance of the tube, and on the other hand, a component which is added by the negative feedback effect of the tube.

Before we can respond to this second part, it is necessary to consider precisely the effect of the neutralization bridge in Figure 2.1 which is drawn separately in Figure 2.2. We do this first with a "cold" tube. The bridge branches are formed by the two sections of the input circuit coil and the two capacitances C_{ga} and C_N. This means that the grid-anode capacitance of the triode C_{ga}, which includes the associated socket and parasitic circuit capacitance and the neutralization capacity C_N, which includes the internal anode-cathode capacitance and the socket and wiring capacitance. We can define the tap ratio for the coil as the voltage

[2]European/German FM-band 88.5MHz-100MHz in 1953

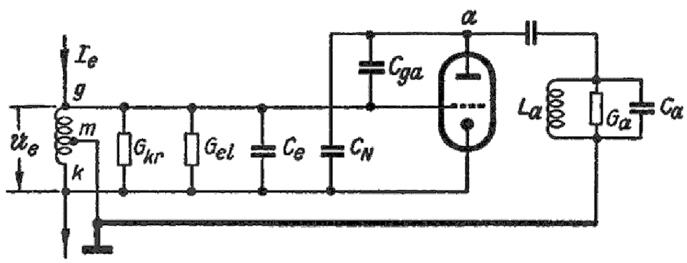

Bild 4. Gegenkopplungsschaltung, schematisch

Figure 2.1: Schematic for negative feedback topology

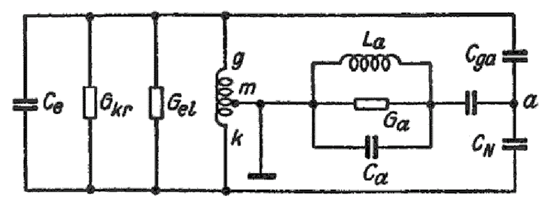

Bild 5. Brückenschema zu Bild 4.

Figure 2.2: Bridge Schematic from Figure 2.1

ratio $x = \frac{U_{mk}}{U_{gk}}$, with U_{mk} as the RF voltage between the points m and k and U_{gk} as the voltage between points g and k. For a resonant circuit, which is not heavily damped, one can apply this simple definition without concern for the stray or leakage flux between the two winding sections. Even the specific construction of the coil is not critical. The same definition also applies to a corresponding capacitive voltage division. The output circuit of the triode stage, consisting of L_a, C_a and G_a, is located in the $m-a$ bridge diagonal, if $\frac{C_N}{C_{ga}} = \frac{1-x}{x}$. The C_{ga} and C_N capacitances affect the resonance tuning of both circuits. The series capacitor $\frac{C_{ga}C_N}{C_{ga}+C_N}$ is added to the input capacitance C_e, while you have to include $C_{ga} + C_N$ in the output side as approximately added in parallel to capacitance C_a.

In "warm" tube, capacitances C_{ga} and C_N affect the tuning in the same way. But they convey also then no feedback from the output to the input, as long as the bridge is balanced. Feedback comes to the input circuit because part of the input current flows through the cathode as current I_k. This is the output circuit at resonance

$$I_k = I_a = \frac{S \cdot U_{gk}}{1 + \frac{DS}{G_a}}; \qquad D = \frac{1}{\mu}$$

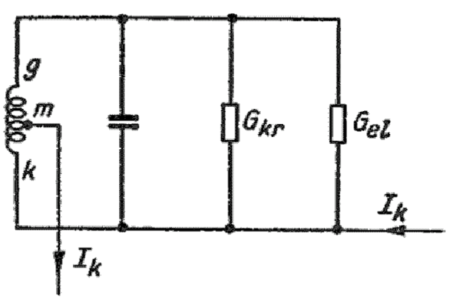

Bild 6. Eingangskreis, vom Katodenstrom durchflossen

Figure 2.3: Input circuit with cathode current I_k flow

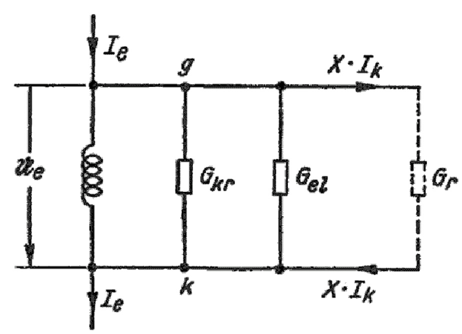

Bild 7. Eingangskreis mit Ersatzstrom

Figure 2.4: Input circuit with equivalent current I_e

According to the schematic in Figure 2.3, I_k flows into the input circuit at k and comes out the tap at m. If we are to base the definition on the x tap ratio, we may equate the effect of this current with respect to the overall effect of an input circuit current $x \cdot I_k$ which enters at k and exits at g. This affects feedback current, as illustrated by Figure 2.4, as if part of I_e it were consumed as $x \cdot I_k$, in a circuit in parallel to the underlying conductance $G_r = \frac{x \cdot S}{1+\frac{DS}{G_a}}$. Additionally, it is noted that for frequencies which differ by $\Delta\omega$ from the resonant frequency of the output circuit, the following applies:

$$G_r = \frac{x \cdot S}{1 + \frac{DS}{G_a + j \cdot 2\Delta\omega(C_a + C_{ga} + C_N)}}$$

This means both an increase in the real component and the addition of an imaginary component in the feedback conductance.

Another imaginary component is added, if the effect of transit time on Transconductance may no longer be regarded as a strictly real component. In a modern UHF triode, such as the EC92 and ECC81, the phase angle in the 100 MHz band is still quite small (about 8^0), so that one hardly needs to account for it. Additionally, in the 200 MHz band, the phase angle of the Transconductance has no significant effect in the real component of input admittance, while the imaginary component can be compensated by retuning.

3 Gain calculation with power matching, bandwidth

The Calculation of RF amplification stage gain with an input transformer depends on the ratio of the transformer and the voltage gain from the tube itself. For an FM-band radio receiver, it is useful to specify the preamp gain from an antenna input with $R_A = 240\Omega$ to the grid of the mixing tube. If the latter is connected directly at the hot end of the output circuit, then

$$A_{\max} = \ddot{u} \cdot \frac{S}{G_a + DS}$$

The same formula can also be used with any coupling mode of the mixer, if one understands the conductance G_a of the coupling element (see below, section 5!). Power matching must be done with $G'_A = G_{kr} + G_{el} + \frac{xS}{1+\frac{DS}{G_a}}$. It then applies to the input transformation ratio

$$\ddot{u} = \frac{1}{\sqrt{R_A \left(G_{kr} + G_{el} + \frac{xS}{1+\frac{DS}{G_a}} \right)}}$$

and gain value

$$A_{\max} = \frac{S}{(G_a + SD)\sqrt{R_A \left(G_{kr} + G_{el} + \frac{xS}{1+\frac{DS}{G_a}} \right)}}$$

You can now set the gain for a particular circuit and tube data with the calculation as a function of x as the tap ratio. Figure 3.1 has such a curve. The curve shows $A_{\max} = A_{\max}(x)$ for the EC92 single triode or the ECC81 dual triode. It was based on the following values:

$G_{Kr} = 0,1\ mS$	$G_{el} = 0,05\ mS$	$R_A = 0,24\ k\Omega$
$S = 5\ mS$	$G_a = 0,5\ mS$	$D = 0,02$

$mS =$ Mili-Siemens $(k\Omega^{-1})$.

The somewhat low value for G_{el} is for the purely electronic component and the damping contribution of the cold tube. The usual measurement of G_{el} gives higher values, in which is already included a certain feedback conductance which arises from the finite inductance at the cathode, that is, such measurements are in fact not executed with $x = 0$.

The sweep of the entire input conductance

$$G_e = G_{Kr} + G_{el} + \frac{xS}{1 + \frac{DS}{G_a}}$$

is also plotted in Figure 3.1:[1] it increases linearly with increasing x. The same curve also represents the input circuit bandwidth with a corresponding frequency scale. The antenna is not included in this bandwidth plot.

[1]In Fig. 3.1 the gain (in German: "Verstärkung") is denoted as $V_{\max}$, V_{opt}. For convenience, in the text the gain is denoted $A_{\max}$, A_{opt}.

By connecting the antenna, then the bandwidth is increased even more noticeably, such that the performance adjustment to an antenna structure with approximately the same intrinsic bandwidth is 1.4 times. It must be noted that the bandwidth information contained in this configuration assumes that G_a can be considered constant. With a tuned output circuit, this would obviously not be the case while the test frequency is changed.

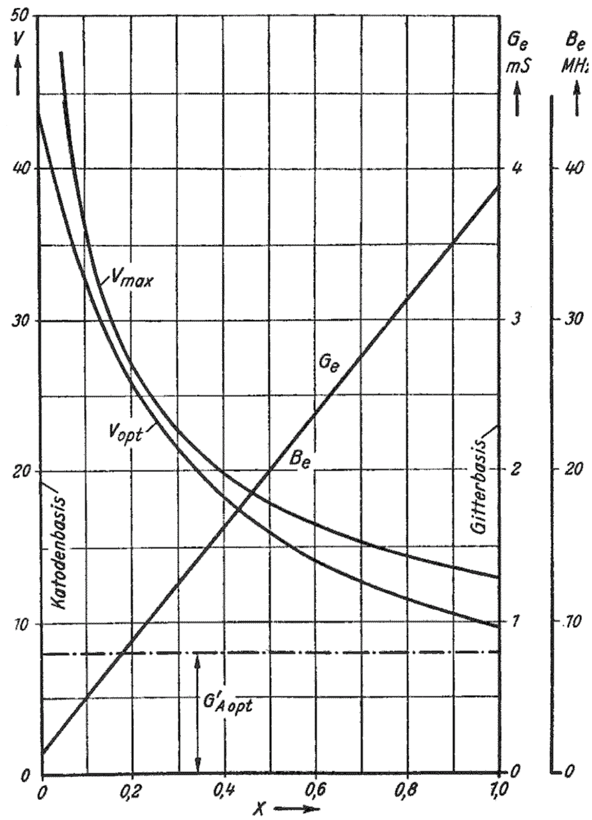

Bild 8. Eingangswirkleitwert, Eingangskreis-Bandbreite für 16 pF gesamte Kreiskapazität, Verstärkung V_{max} bei Leistungsanpassung und Verstärkung V_{opt} bei Rauschanpassung

Figure 3.1: Input effective conductance, input circuit bandwidth for 16 pF total circuit capacitance, $V_{\max} = A_{\max}$ optimum power gain (**OP**) and $V_{\text{opt}} = A_{\text{opt}}$ Noise-Optimal gain (**ON**)

A relatively narrow band output circuit would then cause a resonance curve in the input voltage U_e according to Figure 3.2.

These test conditions do not apply in practice. You will generally have a "continuously tuned" circuit in the output of the RF stage in the FM-band radio receiver. By setting the correct synchronism with the oscillator, then changes of the value of G_a for each receiving frequency within the FM-band broadcast band are insignificant.

For a fixed mid-band tuned output circuit, for example, in television receivers, the modification of G_a is also limited because this circuit must have a large bandwidth.

4 Optimization for the limiting sensitivity (lowest noise)

It was mentioned already that you can reach the highest limiting sensitivity with a specific antenna match which is independent of x. $G'_{A_{\text{opt}}}$ denotes the necessary transformed antenna conductance for "**ON**". It has

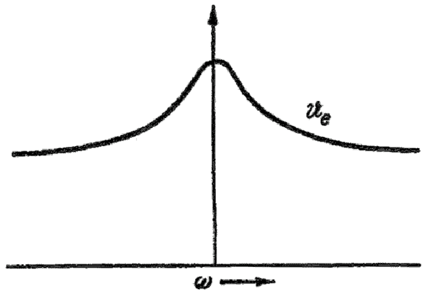

Bild 9. Resonanzkurve am Eingangskreis, wie sie sich bei feststehender Abstimmung eines schmalbandigen Ausgangskreises ergeben würde

Figure 3.2: Resonance curve at the input circuit, as it would result from fixed tuning of a narrow-band output circuit

also often been found that the noise optimum is not exactly at the resonant tuning of the input circuit, but at a certain small capacitive detuning of the same. This is expressed by $j \cdot g_e$ imaginary susceptance in parallel with G_{kr}. These ratios for the noise optimization were the subject of several investigations. A new work by H. Rothe [3] should be pointed out, where you can find the other authoritative literature also also included. The noise optimization formulas have also been given in these works. While the older formulas have ignored certain important factors that are given in newer formulas, they give a deeper insight into the physical relationships, especially Rothe's formulas. Discussing these formulas would lead us outside the present scope.

The following study optimized the best limiting sensitivity for the EC92 tube with $G_{kr} = 0,1\ mS$ determined by measuring: $G'_{A_{\text{opt}}} = 0,8\ mS, g_e = 0,45\ mS$ for $f = 93\ MHz$ (center of the FM broadcast band $[88.5 - 100$ MHz]).

In Figure 3.1 has a horizontal line drawn according to this value $G'_{A_{\text{opt}}} = 0.8\ mS$. Apart from the aforementioned, the **OP** and **ON** optimizations meet where this horizontal straight line intersects the rising G_e line. The example was chosen for the case where $x = 0,17$.

The gain in the noise-optimal setting A_{opt} is always smaller than the gain with performance optimization $A_{\max}$. The quotient $A_{\text{opt}}/A_{\max}$ can be calculated with reference to Figure 4.1 easily.

Bild 10. Einströmungsschema bei kapazitiv verstimmtem Eingangskreis

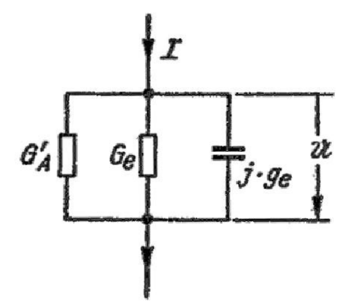

Figure 4.1: Input current schematic with input circuit capacitively detuned

We set a I current in the parallel circuit. With $G'_A = G_e$ and $g_e = 0$, we get a voltage of $U_{\max} = \frac{I}{2G_e}$. This yields power matching (gain optimization), I.e. the value of the power dissipated at the source equals the value of the power consumed at the load:

$$N_{\max} = \frac{I^2}{4G_e} = \frac{I^2}{4G'_A}$$

For the comparison of the gains at different values of G'_A und g_e a constant value of the available source power $N_{\max}$ must be assumed, i.e. the input current must be set according to

$$I = 2\sqrt{G'_A \cdot N_{\max}}.$$

for each value of the inflow G'_A.

Since we now want to compare U with $U_{\max}$ for arbitrary values of G'_A and g_e where I flows, we use the relationship

$$N_{\max} = U_{\max}^2 \cdot G_e$$

and taking into consideration detuning, we get:

$$U = \frac{I}{G'_A + G_e + j \cdot g_e} = \frac{2\sqrt{G'_A} \cdot \sqrt{G_e} \cdot U_{\max}}{G'_A + G_e + j \cdot g_e}$$

from which we obtain after some rearrangement:

$$\left| \frac{U}{U_{\max}} \right| = \frac{2}{\sqrt{\frac{G'_A}{G_e} + 2 + \frac{G_e}{G'_A} + \frac{g_e^2}{G'_A \cdot G_e}}}$$

According to this formula, a curve for the quotient $A_{\text{opt}}/A_{\max} = |U/U_{\max}|$ was calculated for the earlier data (Fig. 4.2). A_{opt} was also shown in Figure 3.1. It is seen that A_{opt} and $A_{\max}$ for both $x = 0$ (**GC**-circuit) and for $x = 1$ (**GG** circuit) are much lower than in an intermediate region in the vicinity of $x = 0.2$.

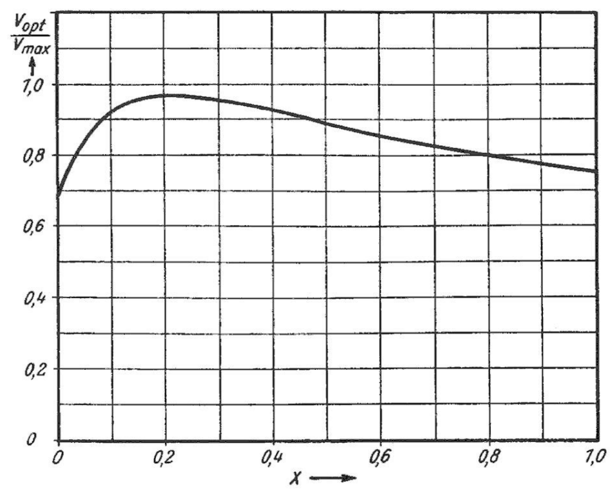

Bild 11. Verhältnis der Verstärkungszahlen V_{opt}/V_{max} in Abhängigkeit von x

Figure 4.2: Ratio of the gain values $A_{\text{opt}}/A_{\max}$ as a function of x

5 Practical considerations for the choice of the x tap ratio

The relatively flat shape of the curve for $A_{\text{opt}}/A_{\max}$ in Figure 4.2 gives a certain latitude for choosing the best x-value, so that other factors may be taken into account. These factors include gain, bandwidth, the required neutralization capacity and stability against neutralization errors.

If we design a circuit with the calculations above, then the maximum value for $A_{\text{opt}}/A_{\max}$ is for $x = 0.2$. In this case, the required neutralization capacity is 10 pF, if we estimate C_{ga} at 2.5 pF. This gives a bandwidth of 8.5 MHz and a gain $A_{\text{opt}} = 26$ which is 2.7 times larger than that achievable with a grounded grid circuit. One can perform a thoroughly reliable neutralization at this gain level. However, it has been found that a capacitive voltage divider for the input circuit in Figure 5.1 is better than a tap at the input circuit coil.

A better bridge neutralization minimum is obtainable with the capacitive divider. On the other hand, the risk of wild oscillations with a high gm triode in the decimeter wavelength region is also greater with a capacitive divider. The tendency for these oscillations can be suppressed best by a combination of a small

Bild 12. Gegenkopplungsschaltung mit kapazitiver Spannungsteilung am Eingangskreis und Dezimeter-Schwingschutz in der Katodenleitung

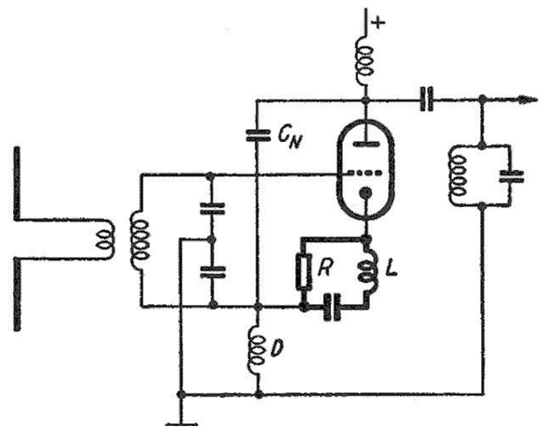

Figure 5.1: Negative feedback circuit with a capacitive voltage divider at the input circuit and decimeter wave oscillation protection in the cathode path

self-inductance (20 nH) with a parallel effective damping resistance (60 to 200 Ω, tenth of a Watt size) and a cathode bypass capacitor (100 pF) which are in the cathode path, as is also shown in Figure 5.1. The small self-induction has an extraordinarily strong negative feedback effect for decimeter wavelength oscillations, but at the fundamental frequency, it only has a small effect on the x-value which, by appropriate choice of the capacitive divider ratio, can be corrected again.

The design just discussed is chosen for the front end stage of a broadband FM-band radio receiver, and the front end is fixed-tuned at mid-band which is a favorable design. You can, however, without worsening significantly the relation $A_{\text{opt}}/A_{\max}$, choose a smaller x-value and thus provide significantly higher amplification. This is a likely consideration for a high-end FM-band receiver with three-gang tuning. For example, $V_{\text{opt}} = 27.4$ is achieved with $x = 0.1$ which is more than three times higher gain than with a pure grounded grid circuit. The bandwidth is 5 MHz and is no longer sufficient for a good alignment of the input circuit in the center of the 100 MHz FM-band band. The neutralization at this high gain is quite critical, and the required neutralization capacity would have to be 20pF in the previously specified circuits. Such a large value is unsuitable for many purposes, because C_N appears as a parallel capacitance to the output circuit and the total capacitance of this circuit is thus inadmissibly enlarged.

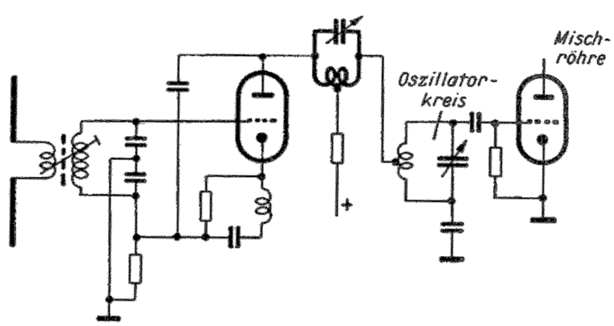

Bild 13. Gegengekoppelte Vorstufe mit zwischen Vorstufenanode und Eingangspunkt der Mischstufe liegendem Ausgangskreis

Figure 5.2: Fedback input stage with output circuit connected between its anode and the entry point of the mixer circuit.

You can avoid this difficulty in two ways. One approach is to use a balanced output circuit (Figure 5.2). This balanced circuit is not between anode and ground, but between the anode of the preamp and the grid of the mixer, or the point of symmetry at the grid tuned circuit of the mixer, while the point of symmetry at the output circuit is connected to ground. The approximately symmetrical distribution is then performed by the roughly equal capacitances of the two hot points of the circuit to ground which are given on the anode side by $C_{\text{ga}} + C_N$, and by the input capacitance of the mixer circuit on the mixer side. Such a circuit can be configured, for example with a three-section tuning capacitor with insulated tuning sections.

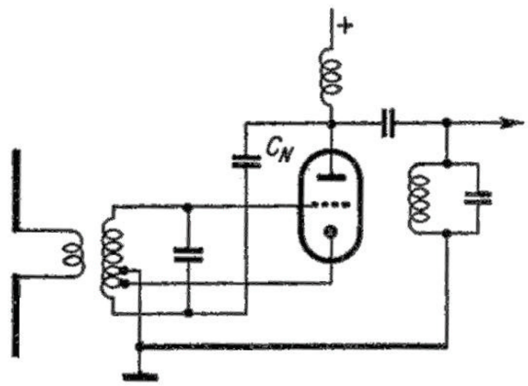

Bild 14. Gegenkopplungsschaltung mit zwei Anzapfungen der Eingangskreisspule

Figure 5.3: Negative feedback circuit with two taps of the input circuit coil

The second approach circumvents the need for a large capacity for C_N in that it provides two taps at the input circuit, as shown in figure 5.3. The cathode of the tube is no longer at the "lower" end of the input circuit, but on the second tap. The first tap tied to ground then sets the symmetry point of the entire circuit. It is then $C_N = C_{\text{ga}}$. A corresponding circuit with a capacitive voltage divider is shown in Figure 5.4. You can see a parallel combination of a self-inductance L and resistor R in at the output circuit, which has the task of suppressing decimeter wavelength oscillations.

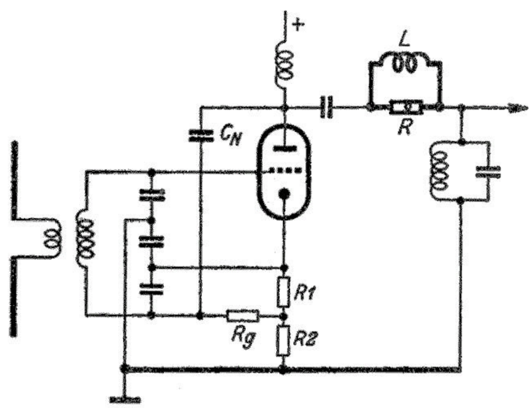

Bild 15. Gegenkopplungsschaltung mit zweimal kapazitiv aufgeteiltem Eingangskreis sowie mit Dezimeter-Schwingschutz in der Zuleitung zum Ausgangskreis

Figure 5.4: Negative feedback schematic for capacitively split input circuit with two taps and with decimeter wave oscillation protection in the output circuit connection.

While you can get especially high gain figures with the aforementioned design, there is of course also the opposite possibility, with the tap in the direction of larger x values (i. e. $x = 0.5$). Such a calculation results in a particularly wide bandwidth and insensitivity to neutralization inaccuracies. Choosing just $x = 0.5$, one obtains again a balanced input circuit. You can replace these lumped inductances and capacitances with appropriately sized pieces of FM-band twin-lead. One can, for example, go from the two antenna terminals with such a piece of line to the grid and cathode of the tube, and also connect to the antenna terminals a second twin-line section, whose other ends connect high frequencies to ground. By suitable choice of the lengths of these line segments you can achieve both the right tuning and the proper adjustment of the antenna resistance. This arrangement has the advantage of extreme simplicity, since it requires no special means to suppress decimeter wavelength oscillations.

The new negative feedback circuit, has an advantage over the conventional **GC** circuit, especially in relation to the stability of the neutralization. This can be demonstrated on the basis of the two circuits in figures 5.5 and 5.6.

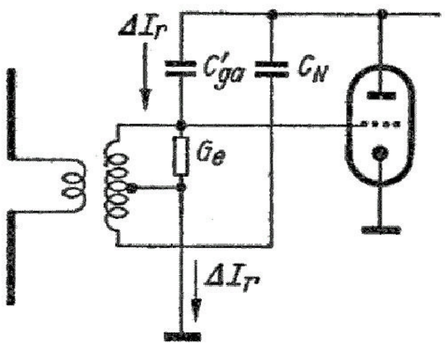

Bild 16. Rückkopplungseinströmung bei ungenauer Neutralisation in den Eingangskreis einer KB-Schaltung

Figure 5.5: Feedback current flow with inaccurate neutralization in the input circuit of **CC** topology.

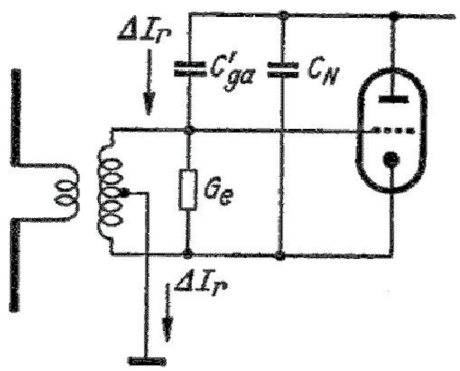

Bild 17. Rückkopplungseinströmung bei ungenauer Neutralisation in den Eingangskreis der Gegenkopplungsschaltung

Figure 5.6: Feedback current flow with inaccurate neutralization in the input circuit of the negative feedback loop topology

For a specific neutralization error, it is outweighed by the influence of the grid-anode capacity: $C'_{\text{ga}} = C_{\text{ga0}} + \Delta C_{\text{ga}}$. This gives a feedback current into the input circuit with the value $\Delta I_r = j\omega\Delta C_{\text{ga}} \cdot U_a$. The total conductance between grid and cathode in both circuits is $G_e + G'_A$. In the case of the **GC**-circuit (Figure 5.5) a feedback voltage is then obtained between grid and cathode as

$$U_r = \frac{\Delta I_r}{G_e + G'_A}$$

while this voltage in the circuit of Figure 5.6 is only

$$U_r = \frac{I_r}{G_e + G'_A} \cdot (1 - x)$$

. In the case for $x = 0.5$, they will be under otherwise identical neutralization error conditions, but the destabilization effect is only half as great as with a **GC**-circuit with the same gain characteristics. The negative feedback circuit thus tolerates twice as large a neutralization error than the corresponding **GC**-circuit.

6 Measurements to verify the theory

The validity of the formula for G_e as a function of x, was verified first. No FM-band structure is known to determine easily the x tap ratio with the accuracy required for such a verification. A "model circuit" was therefore built with a tapped transformer that was measured exactly at a frequency of 100 kHz (Figure 6.1).

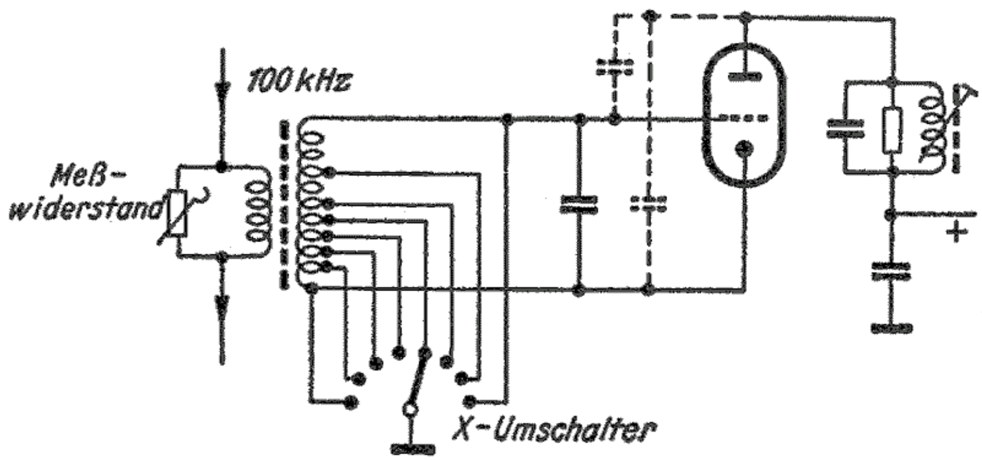

Bild 18. „Modellschaltung" für grundsätzliche Messungen mit 100 kHz Meßfrequenz

Figure 6.1: "Circuit simulation model" for basic measurements with 100 kHz measurement frequency.

The data were

$$\begin{array}{lll} G_{Kr} = 0,11\ mS & G_{el} = 0 & R_A = 0,6\ k\Omega \\ S \ = 5\ mS & G_a \ = 0,2\ mS & D \ = 0,02 \end{array}$$

The winding of the input circuit had taps for $x = 0.1;\ 0.2;\ 0.3;\ 0.4;\ 0.5;\ 0.7;\ 1.0$. The input circuit was now driven by a 100-kHz generator via a "current coupling capacitor", which was so small that you could count on it for a constant current inflow, regardless of the circuit loading. A calibrated resistor was connected in parallel with the primary winding of the input transformer and its voltage drop was measured at the input circuit. We compared the input currents which resulted with and without the parallel resistor for the same voltage at the input circuit. The parallel resistance, which required double the current flow, was then the same size as the primary-side input resistance under measurement. The results of such measurements at different x-values, converted to the secondary side, are shown as points in Figure 6.2 along with the curve for the calculated resistance values from the formulas above.

These measurements were performed first so that the circuit contains no capacitance, which corresponds to the capacity of C_{ga} and C_N during FM-band operation. At the low operating frequency of the circuit model a neutralization at the selected design values was indeed not required. Now it was verified if the conditions change when such capacities are represented. Thus, additional capacity for C_{ga} and C_N is applied in each corresponding ratio for the variable x (shown by the above relationship $\frac{C_N}{C_{\text{ga}}} = \frac{1-x}{x}$). During the inspection of the resonance points of the input and output circuit, it appeared that the two capacitors affect tuning in the manner described in Section 2. With these capacitances, for example, which amounted to 1000 pF for $x = 0.5$, a measurement of the primary-side input resistance was also performed; there being no significant deviations from the measured values without additional capacity. The experimental verification confirmed the accuracy of the considerations in Section 2, namely, that the neutralization bridge in proper alignment does not affecting the input resistance. This was confirmed also by a calculation carried out by H. Schubert and the author, for the input conductivity G_e with respect to these capacitances. This more accurate calculation also showed no significant differences compared to the G_e values calculated under Section 2.

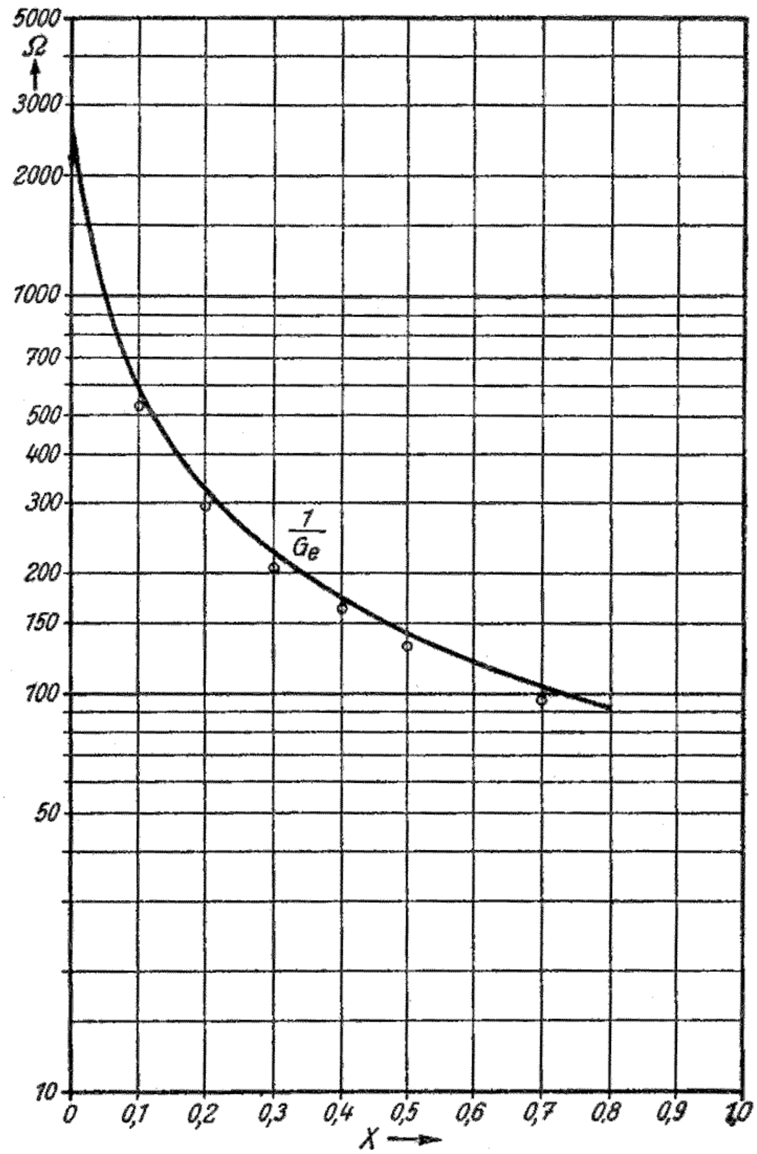

Bild 19. Errechnete und mit der Modellschaltung gemessene Werte von $1/G_e$ (Ordinate logarithmisch). Ausgezogene Kurve: gerechnet, einzelne Punkte: gemessen

Figure 6.2: Calculated and measured $1/G_e$ values of the model circuit (logarithmic ordinate). Solid curve: calculated, individual points: measured.

Practical steps to design an RF stage for an FM-band receiver with an EC92 single or an ECC81 dual triode yielded gains from 20 to 30 from a 240-ohm antenna. These values, which were obtained for $x = 0,3 \cdots 0,4$, agree also quite well with the expectations. To test the stability of such a circuit, neutralization was been calibrated once and 15 different samples of ECC81 tubes were tested. The gain values changed only slightly from tube to tube (usually below 10%). A measurement of the limiting sensitivity resulted in a noise figure of about 3.0 [9.5dB] in the 100 MHz band, (defined as overall noise figure). A comparison with a correspondingly constructed grounded cathode circuit confirmed the expected theoretical noise figures.

References

[1] Die Telefunken-Röhre, Sonderheft zum 70. Geburtstag Prof. Dr. Dr. Ing. e.h. Hans Rukop, herausgegeben von Dr. Ing. Horst Rothe, Leiter der Röhrenentwicklung TELEFUNKEN, Franzis-Verlag München, 1953, Hochfrequenzverstärkung mit Trioden, R. Cantz, pp. 52 – 69

[2] Rothe, H.: *Die Empfindlichkeit von Empfängerröhren*, AEÜ 3 (1949), 233 – 240.

[3] Rothe.H.: *Die Grenzempfindlichkeit gittergesteuerter Röhren*, Tfk-Röhre H. 30 (Sonderheft 1953), 7 – 21.