Frequency compensated LC networks for oscillators with the wide tuning range.

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Introduction.

Anyone who have designed a capacitively tuned VFO/VCO with the wide tuning range (>1.5-2x) has noticed that the oscillation threshold changes substantially with the tuning frequency. The usual solution to the problem is a combination of:

1. the AGC circuit, most frequently via the amplifier bias that changes the amplifier conduction angle;
2. shunting the LC circuit with a damping resistor or by a low impedance amplifier terminal;

Both methods have a common drawback - amplifier gain must be increased to maintain the oscillation condition margin on both sides of the tuning range and/or to compensate for the damping resistor losses. Increasing gain requires either more complex amplifier or tighter coupling of the resonant circuit to the amplifier. The later increases the influence of amplifier parasitic elements on the resonant circuit. The more complex amplifier can increase the phase jitter. Both solutions have a negative impact on the oscillator frequency stability as well as the phase noise.

There is a passive solution to the problem that requires neither damping the Q-factor of the resonant circuit nor a large dynamic range AGC. As far as I know it first appeared in the research paper by Vackar [1] published back in 1949. Vackar is best known for his oscillator topology shown on fig.5 in [1] (see below). However it is the "extended range" Vackar oscillator shown on fig.6 that introduces the first oscillator arrangement with nearly flat oscillation threshold across the wide tuning range:

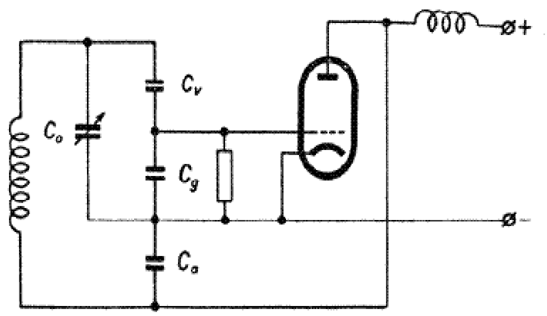


Fig. 5

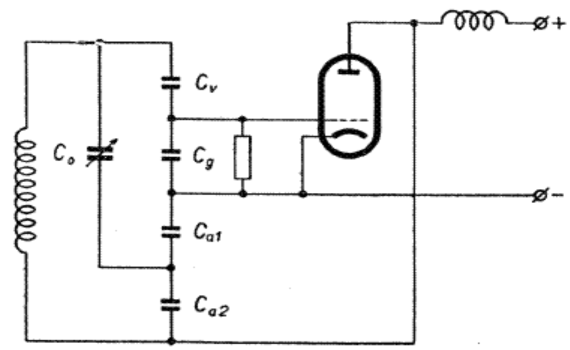


Fig. 6

Somehow, despite simplicity, this idea eluded the radio engineers attention and was forgotten. No modern textbook or oscillator design reference material that I know of cover either the "extended range" Vackar circuit or the principles behind the LC networks with frequency compensated oscillation threshold.

After building few LC oscillators and facing the usual tradeoff I decided to do a simulation study of various oscillator topologies in LTSpice. By that time I already had access to the original Vackar paper thanks to the invaluable efforts of N1EKV. The main result of this study was realization that there exists a whole class of oscillator topologies with properties similar to the "extended range" Vackar with respect to the frequency compensation of the oscillation threshold. These results encouraged me to:

1. build and verify the concept in the several hardware prototypes;
2. search for modern or historical references covering this idea;

Test prototypes confirmed the simulation, both results are presented in this paper.

However the search for historical references is still in progress. A similar concept is used in the "padder feedback" oscillator-converter described in RDH4 [2], although the dual function of the padder capacitor (superhet tracking) limits the frequency dependence optimization. A similar scheme (discussed in the simulation results) is briefly mentioned in the technical reference [3] (in Russian, published in 1955).

A little bit of theory.

(In this paper I will focus mostly on general discussion of the simulation results and the prototype measurements. For the mathematical treatment of the circuit operation please refer to a separate theory write-up.)

Any oscillator contains a resonant circuit with frequency dependent energy losses (which can be described as Q(f) = E.stored(f)/E.loss(f) per cycle (note Q factor dependance on the frequency f), and a frequency dependent feedback gain provided by the amplifier (let's call it P(f) = E.added(f)/E.stored(f) per cycle). If there was no frequency dependence of P and Q then the stable oscillation condition is basically Q*P=1 hence it is relatively easy to build a fixed frequency (or narrow tuning range) oscillator using any topology. However in a physical circuit both Q(f) and P(f) have rather complex and different behavior in the frequency domain. This makes it hard to maintain Q(f)*P(f) close to unity over a wide frequency range and requires heavy reliance on the AGC in the feedback loop or various lossy damping methods.

The losses in a simple LC circuit can be separated in two physical components:
1. Rs - series resistance determined by the conductor size/shape material, skin effect, etc.;
2. Rp - parallel resistance determined by the inductor and circuit radiation losses, inductor core losses, amplifier loading, etc.

Both types of losses depend on frequency, so it is more accurate to use Rs(f) and Rp(f). Moreover it can be argued that losses can have some non-linear component and hard to estimate correctly (except maybe for some special experimental setups). However it can be assumed that both Rs and Rp are simultaneously present in any practical implementation.

In computer science and mathematics there are algorithms and processes for which direct mathematical description is hard or even impossible. In such case upper and lower bounds for the metric of interest are proven instead and it is said that the algorithm/process will perform no worse than e.g. the analytically simple algorithm A and no better than the analytically simple algorithm B which are called bounds. In my simulation analysis I use the same approach. Without specific knowledge of Rs(f) and Rp(f) I study the frequency dependence of oscillation threshold for various oscillator topologies with three loss models:
1. model with losses dominated by Rs;
2. model with losses dominated by Rp;
3. model with losses approximately split between Rs and Rp.

Cases 1 and 2 provide bounds on the oscillator threshold behavior. Any physical circuit realization will behave in between these two cases. The third case is used primarily to illustrate how the new class of oscillators can be optimized for any combination of Rs and Rp induced losses.

In my simulations I found that conventional capacitively tuned LC oscillators can be split into at least three classes with vastly different frequency dependence of the oscillation threshold, given the same Rs/Rp loss model. The classes are:
class 1: Armstrong, Colpitts, Hartley, Seiler;
class 2: Vackar (basic form, fig5 in original Vackar paper [1]);
class 3: Clapp/Gouriet.
Within each class the threshold dependence on the frequency is essentially the same. There could be substantial implementation-specific variations, nevertheless when the same loss model is used in the simulation the oscillators within the same class have very similar frequency domain threshold behavior.

None of the oscillators above can be optimized to have flat frequency dependance of the oscillation threshold for an arbitrary mix of Rs and Rp losses. <u>However it is possible to constructively combine two feedback types from two oscillator classes in a single circuit to eliminate at least the linear term of the oscillation threshold frequency dependance</u> (I call it "tilt"). Moreover by properly balancing the amount of feedback between the two paths it is possible to zero out the tilt for any combination of Rs and Rp values. The original "extended range" Vackar and the other circuits in this combined (hybrid) feedback class can be adjusted in-place for zero tilt and have a clearly defined optimum - around this optimum the tilt crosses the zero and changes the sign. For any two points in the tuning range the solution is exact. <u>No dissipative losses are added by the hybrid feedback mechanism, the intrinsic Q of the resonant tank is maintained intact.</u>

Because Rs(f) and Rp(f) in a physical circuit have a complex non-linear frequency dependence there are higher degree terms which are left uncompensated. But practical tests have shown that the remaining terms are surprisingly small, especially if the radiation losses are small and a high Q inductor is used.

It should be noted that for the simplified loss models (frequency invariant Rp-dominant and Rs-dominant losses), the characteristic behavior of the threshold with frequency for conventional oscillators can be approximated analytically without resort to complex impedance calculations, simply by using energy balance computations. (I am planning to add mathematical analysis of the threshold later). The simulation approach below is still important, it gives some additional insight into the resonance peak behavior, and allows quick verification of complex circuits with mixed loss models that would otherwise require a significant effort to analyze mathematically.

It is also important to note that inductively tuned oscillators of the same topology and the same Rs/Rp loss model will have completely different shape of the threshold frequency dependence. However the combined feedback tilt compensation should also be applicable to such oscillators.

Assumptions and Simulation methodology.

For the simulations I used LTSpice which is a free SPCIE simulator and schematic capture software developed by Linear Technology.

Running a full oscillator simulation starting from the buildup of an oscillation up to a steady state condition is a rather slow process and not very suitable for testing multiple topologies at multiple frequencies and three loss models. To speedup the process I decided to break the feedback loop at the input of the amplifier and simply perform AC sweep of the amplifier plus the LC network combination.

Typical oscillator amplifier operates as a voltage driven current source. This makes sense since the voltage input minimizes loading of the parallel LC network output, while the current output minimizes loading of the LC circuit input. A FET or a vacuum tube active device is a good example of such an amplifier, a FET cascode or a pentode tube gets even closer. In simulations I replaced the amplifier with an ideal voltage controlled current source (VCCS) with a preset transconductance Gm. The use of VCCS is not required but makes for easier interpretation of the transfer function: voltage-in to voltage-out. Additionally VCCS allows supplying current to floating nodes from the common test signal when necessary. Effects of the amplifier parasitics are ignored in this simulation, only the properties of the resonant LC networks under the linear loss model are analyzed.

Some circuits also include voltage controlled voltage sources (VCVS). These are added to sense the voltage differential between two floating nodes exactly as the amplifier device sees it (consider how the gate and drain are connected in a Hartley or Colpitts circuits) and re-reference this voltage difference to the ground.

Seven frequency sweeps are performed at different tuning frequencies ranging from 1MHz to 4MHz. For each frequency step the value of the tuning capacitor Ct is estimated based on the frequency. The simulation results for all oscillators in the same class are overlaid on the same graph using different colors. Only the magnitude of the frequency response is displayed (simultaneously plotting the phase response clutters the graph too much). I have intentionally avoided exact frequency and gain match between the oscillators in each class so that curves do not overlap completely. Three models of losses (two bounds and one midpoint) are used for each class. In this simulation Rp and Rs are not frequency dependent. The values of Rs and Rp were chosen as follows:

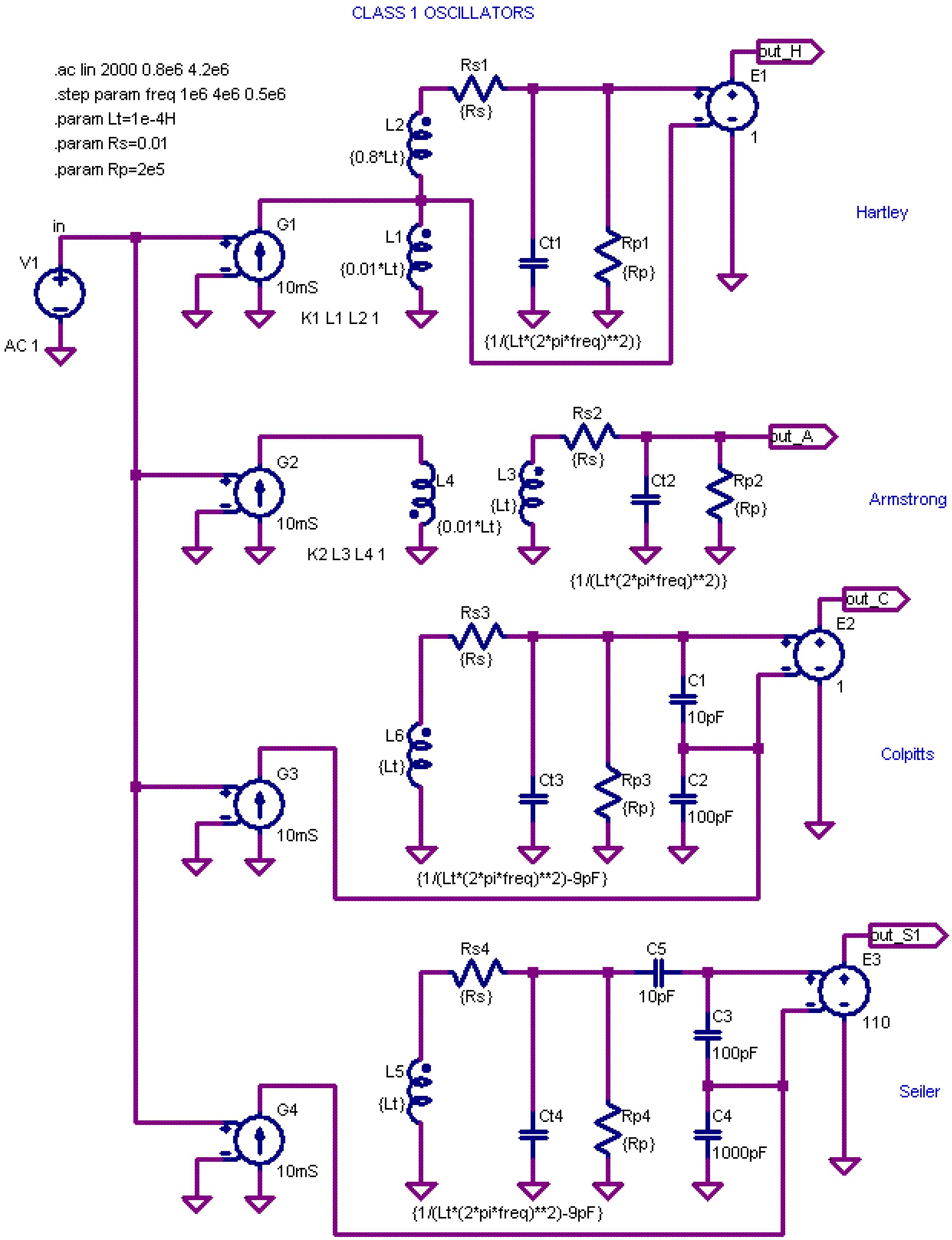
1. Rs-dominant loss model: Rs=10ohm, Rp=1Gohm
2. Rs+Rp midpoint loss model: Rs=10ohm, Rp=200Kohm
3. Rp-dominant loss model: Rs=0.01ohm, Rp=200Kohm

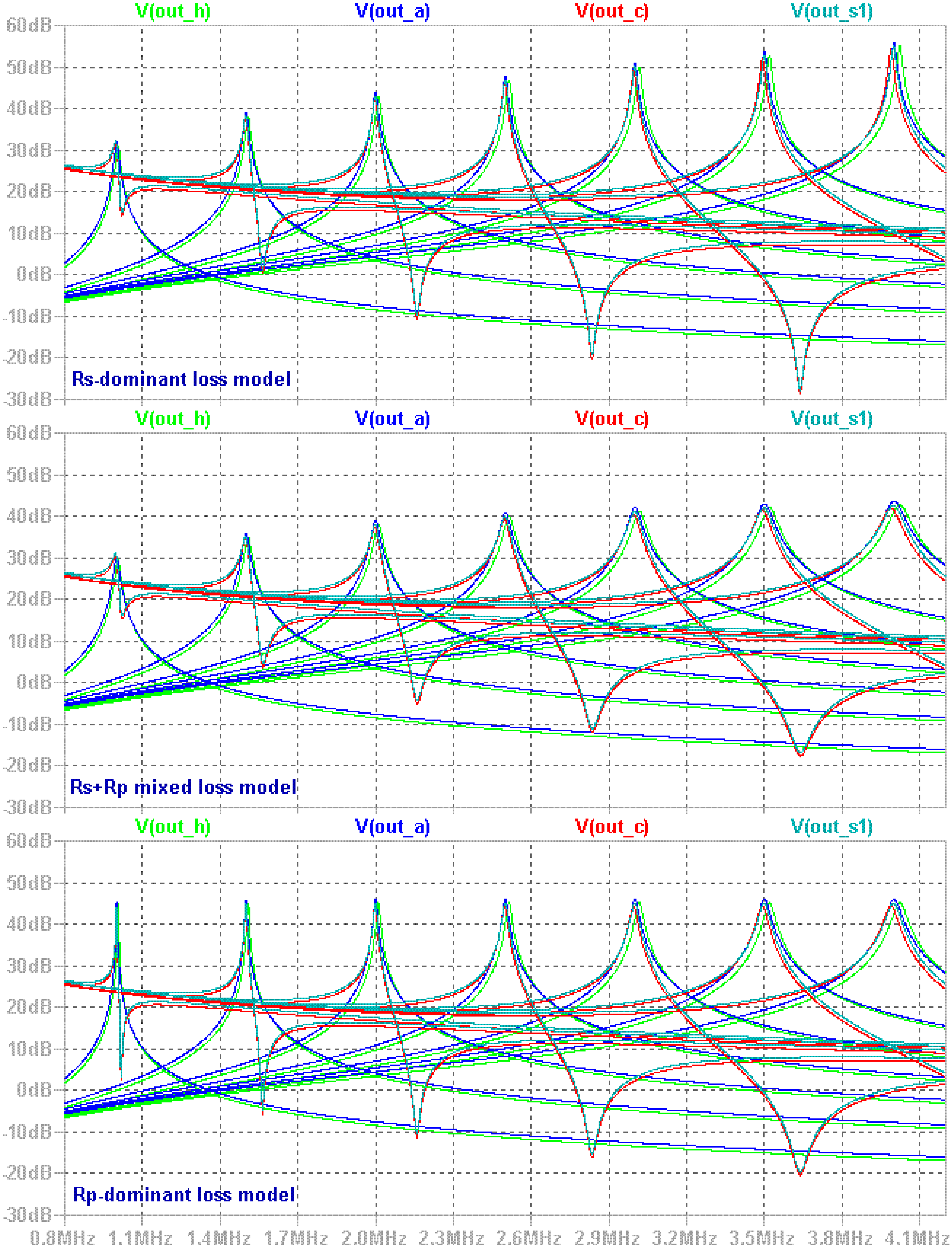
Class 1 oscillators.

Class 1 oscillators are represented by Hartley, Armstrong, Colpitts and Seiler oscillators. The circuit test bench and the simulation results for all class 1 oscillators are shown on the next two pages.

Two notes on the oscillator topology in this class:

1. There is some ambiguity in the literature when it comes to a capacitively tuned Colpitts VFO topology. In this study I use a three capacitor version - main tuning capacitor is connected across the inductor and the divider is comprised of the two additional fixed value capacitors.
2. There is some similarity between the three-capacitor Colpitts oscillator and the Seiler oscillator. The later has additional capacitive divider allowing for more flexible impedance matching to the amplifier parameters. Therefore in the subsequent sections I will use Seiler instead of Colpitts arrangement for one of the hybrid feedback oscillators.


CLASS 1 OSCILLATORS
.ac lin 2000 0.8e6 4.2e6
.step param freq 1e6 4e6 0.5e6
.param Lt=1e-4H
.param Rs=0.01
.param Rp=2e5
in
V1
AC 1
G1
10mS
G2
10mS
G3
10mS
G4
10mS
Rs1
{Rs}
L2
{0.8*Lt}
L1
{0.01*Lt}
K1 L1 L2 1
Ct1
Rp1
{Rp}
{1/(Lt*(2*pi*freq)**2)}
E1
1
out_H
Hartley
Rs2
{Rs}
L3
{Lt}
L4
{0.01*Lt}
K2 L3 L4 1
Ct2
Rp2
{Rp}
{1/(Lt*(2*pi*freq)**2)}
out_A
Armstrong
Rs3
{Rs}
L6
{Lt}
Ct3
Rp3
{Rp}
C1
10pF
C2
100pF
{1/(Lt*(2*pi*freq)**2)-9pF}
E2
1
out_C
Colpitts
Rs4
{Rs}
L5
{Lt}
Ct4
Rp4
{Rp}
C5
10pF
C3
100pF
C4
1000pF
{1/(Lt*(2*pi*freq)**2)-9pF}
E3
110
out_S1
Seiler


V(out_h)
V(out_a)
V(out_c)
V(out_s1)
60dB
50dB
40dB
30dB
20dB
10dB
0dB
-10dB
-20dB
-30dB
Rs-dominant loss model
Rs+Rp mixed loss model
Rp-dominant loss model
0.8MHz
1.1MHz
1.4MHz
1.7MHz
2.0MHz
2.3MHz
2.6MHz
2.9MHz
3.2MHz
3.5MHz
3.8MHz
4.1MHz

V(out_h) represents the frequency response magnitude for the Hartley oscillator;
V(out_a) - Armstrong oscillator;
V(out_c) - Colpitts oscillator;
V(out_s1) - Seiler oscillator;

Notes for the Rs-dominant loss model plot (top):
One can clearly see that the resonance peaks at the seven test frequencies increase in magnitude with the frequency by >20dB over the frequency span from 1MHz to 4 MHz. The graph uses the linear frequency scale and it can be observed that Q at the peaks also rises with the frequency linearly. There is a difference in the shape of the frequency response between the oscillators away from the primary resonance peak but the peaks exhibit almost identical behavior.

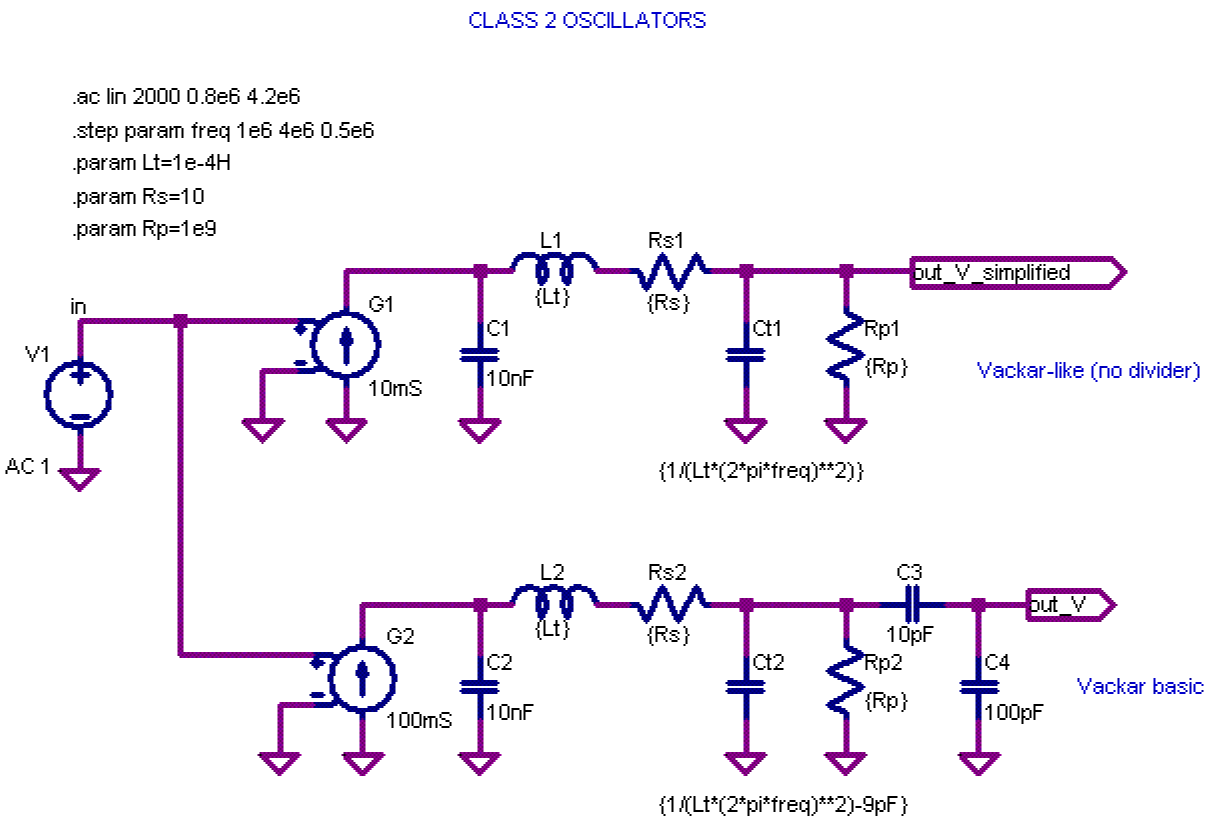
Notes for the Rp-dominant loss model plot (bottom):
When the loss model is dominated by Rp the resonance peaks are almost equal in value. This probably explains why it was a common practice to shunt the tank circuit in this class of oscillators with a relatively small value resistor. A typical grid leak resistor in the tube era was around 50kOhm. With such a heavy loading Rp becomes dominant and the oscillation condition over the tuning range becomes partially equalized.

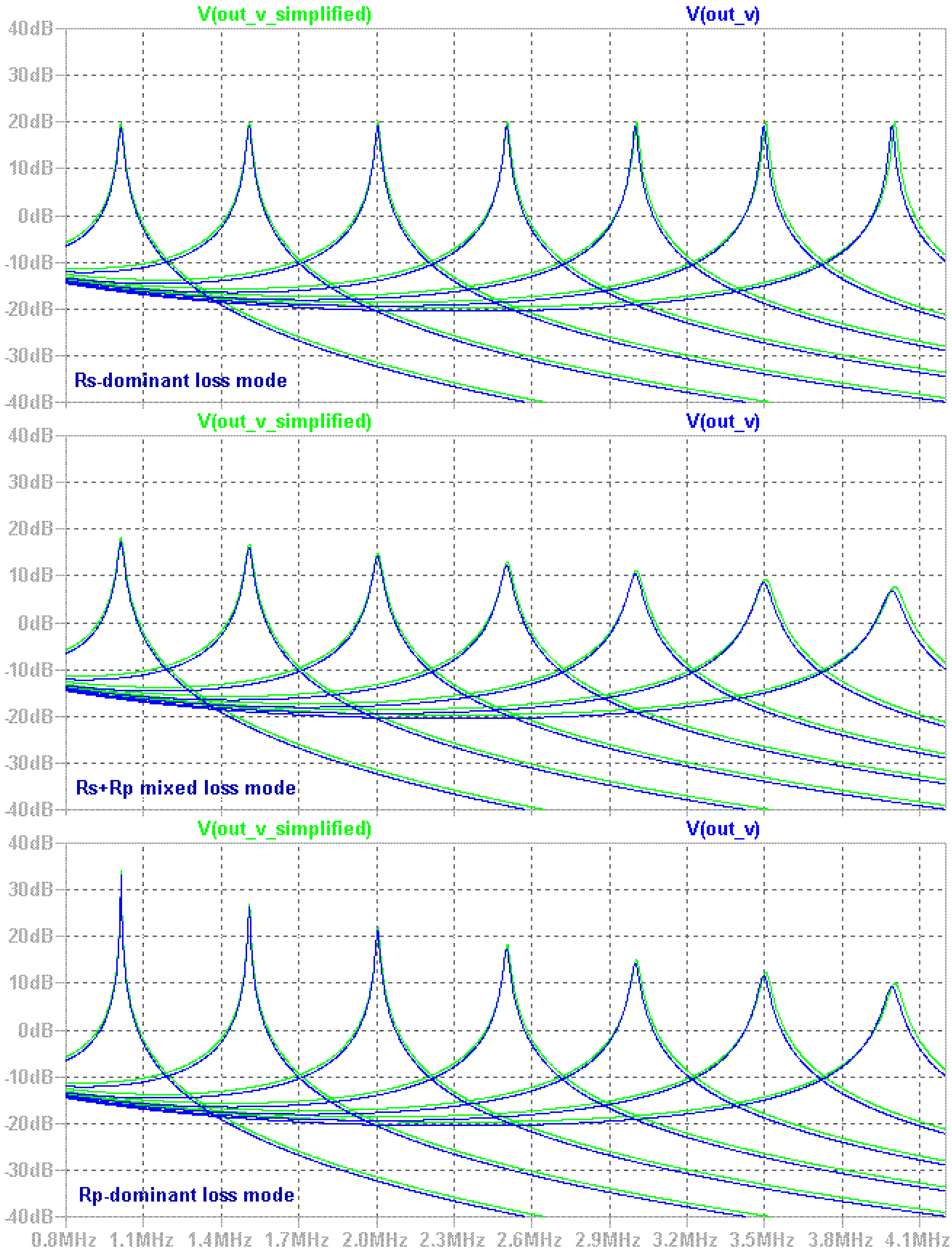
Notes for the Rs+Rp mixed loss model plot (middle):
Not much to say about this picture - the peak magnitude behaves in between the previous two models as expected (there is about 10db magnitude difference between peaks across the tuning range vs >20dB for the Rs-dominant loss model).

Class 2 oscillators.

Class 2 oscillators are represented by the "simplified" and the basic Vackar circuits. Simplified version does not have the divider at the output of the LC network. I included it to emphasize that the divider itself is not critical for shaping of the oscillator frequency response. The role of this divider is different, it: (i) reduces the amplifier coupling to the resonance circuit (minimizing the effects of parasitics and Miller capacitance on the resonance circuit), (ii) allows operating the tank circuit at higher amplitude/energy level (important for the phase noise performance), and (iii) allows easy loop gain adjustment.

The circuit test bench and the simulation results for all class 2 oscillators are shown on the following two pages.




V(out_v_simplified)
V(out_v)
40dB
30dB
20dB
10dB
0dB
-10dB
-20dB
-30dB
-40dB
Rs-dominant loss mode
Rs+Rp mixed loss mode
Rp-dominant loss mode
0.8MHz
1.1MHz
1.4MHz
1.7MHz
2.0MHz
2.3MHz
2.6MHz
2.9MHz
3.2MHz
3.5MHz
3.8MHz
4.1MHz

V(out_v_simplified) represents the frequency response magnitude for the "simplified" Vackar oscillator;
V(out_v) represents the frequency response magnitude for the basic Vackar oscillator;

Notes for the Rs-dominant loss model plot (top):
One can see that the Rs-dominant loss model is a perfect match for the capacitively tuned pi-network used in the Vackar oscillator. Q is rising linearly with the frequency.

Notes for the Rp-dominant loss model plot (bottom):
Now there is >20db tilt in the frequency response. The Rp-dominant loss model is clearly not optimal for the Vackar oscillator.

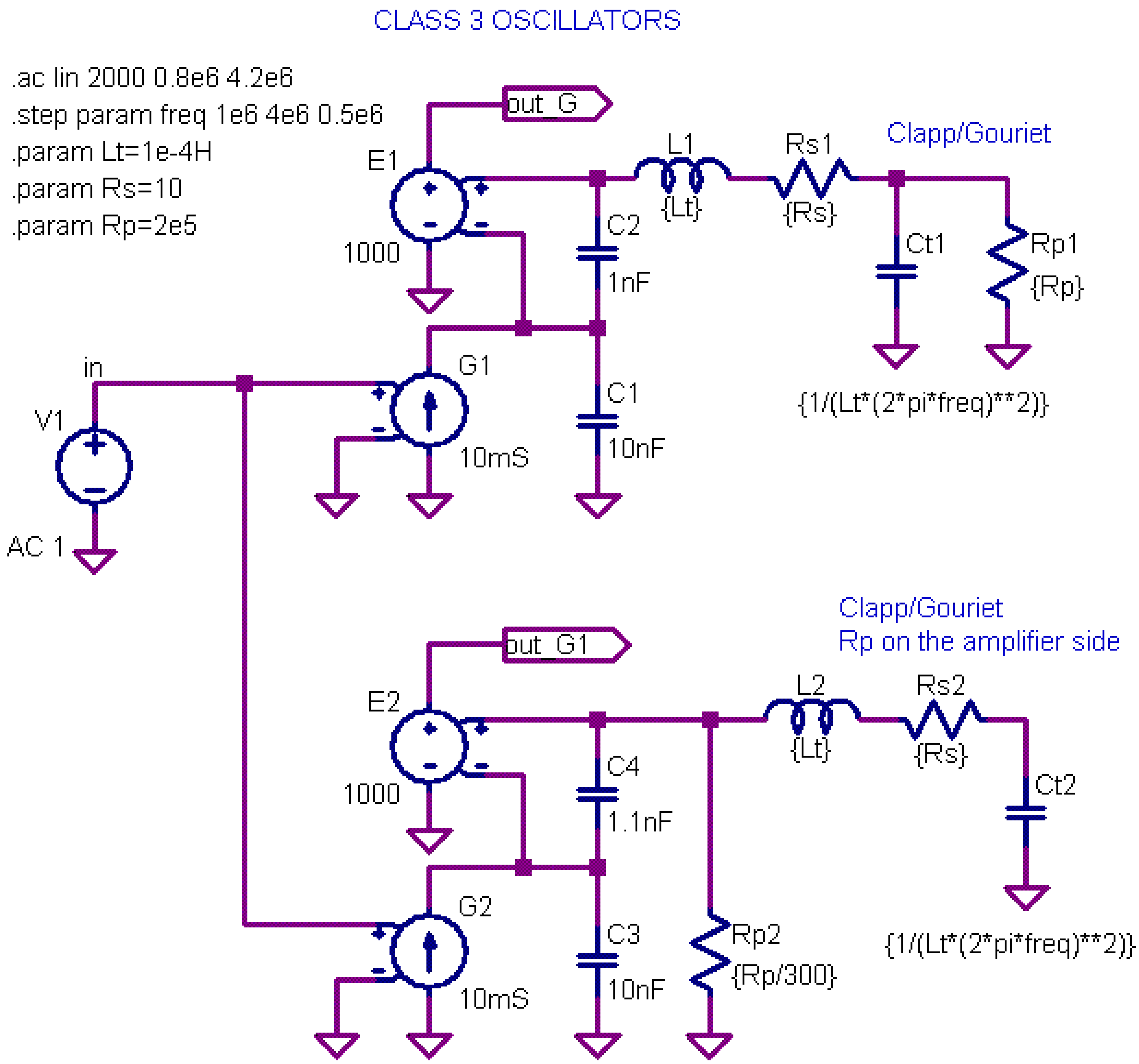
Notes for the Rs+Rp mixed loss model plot (middle):
In a mixed loss model the tilt is smaller, about 10db.

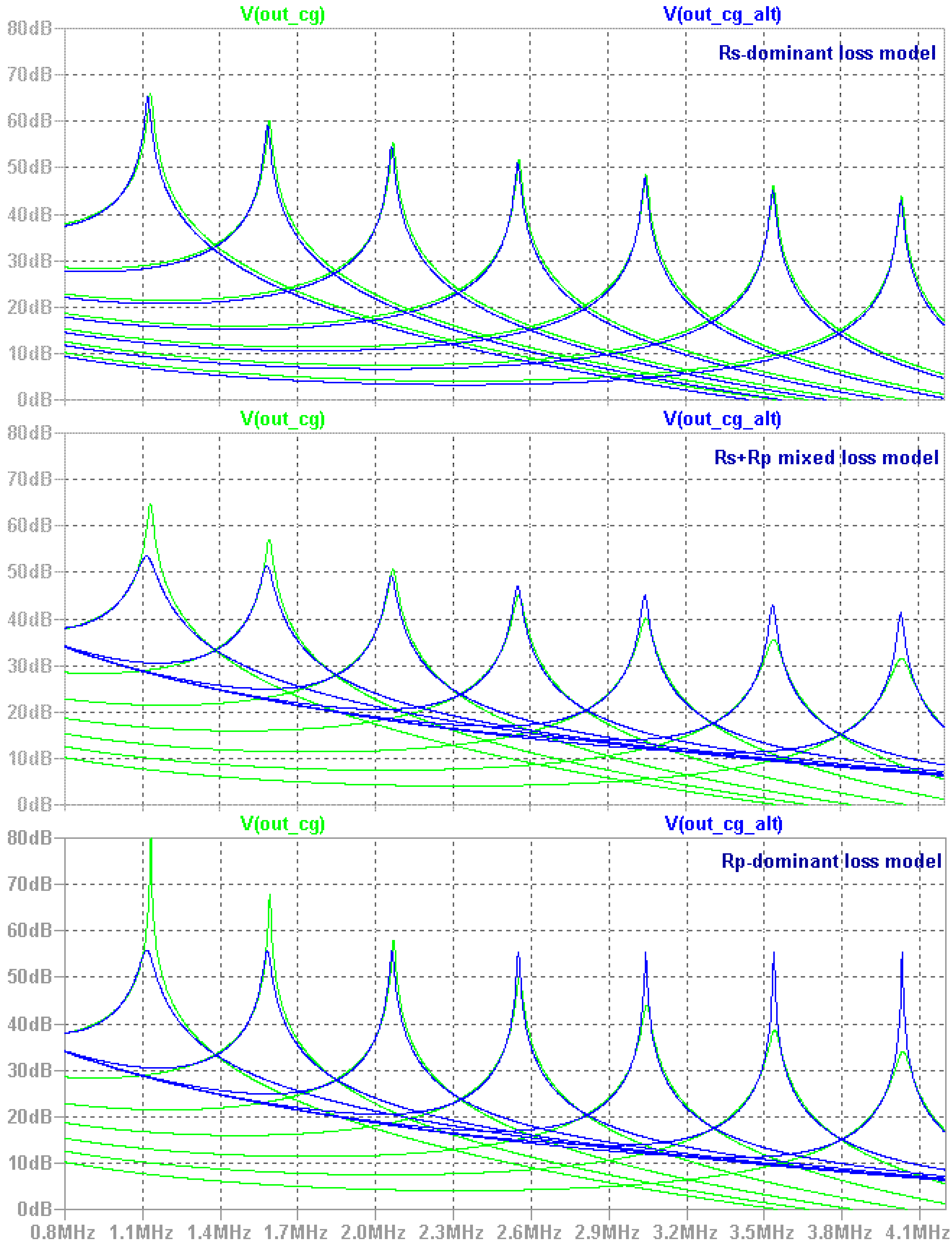
Note that the tilt of the frequency response of the class 2 oscillators is opposite to that of the class 1 oscillators and is comparable in magnitude in all three loss models. The characteristics of the class 1 and the class 2 oscillators are highly complementary. Therefore these two classes are good candidates for the hybrid feedback approach to the tilt compensation for an arbitrary Rp/Rs combination.

Class 3 oscillators.

Class 3 oscillators are represented by the serially tuned Gouriet/Clapp oscillator. I made an exception for this circuit and added a second version where Rp is connected to the amplifier capacitive divider on the side opposite to the tuning capacitor and also Rp has a much smaller value. In case of a heavy amplifier loading (highly unlikely due to large capacitor values typically used in the divider) this version could be more accurate. Nevertheless I think the top circuit much better represents actual losses from multiple factors.

The circuit test bench and the simulation results for all class 3 oscillators are shown on the following two pages.




V(out_cg)
V(out_cg_alt)
Rs-dominant loss model
80dB
70dB
60dB
50dB
40dB
30dB
20dB
10dB
0dB
V(out_cg)
V(out_cg_alt)
Rs+Rp mixed loss model
80dB
70dB
60dB
50dB
40dB
30dB
20dB
10dB
0dB
V(out_cg)
V(out_cg_alt)
Rp-dominant loss model
80dB
70dB
60dB
50dB
40dB
30dB
20dB
10dB
0dB
0.8MHz
1.1MHz
1.4MHz
1.7MHz
2.0MHz
2.3MHz
2.6MHz
2.9MHz
3.2MHz
3.5MHz
3.8MHz
4.1MHz

V(out_cg) represents the frequency response magnitude for the Gouriet/Clapp oscillator;
V(out_cg_alt) represents the frequency response magnitude for the Gouriet/Clapp oscillator with alternative Rp loss model;

Notes for the Rs-dominant loss model plot (top):
There is a 20db backward tilt with the Rs-dominated loss model in this oscillator. Q is rising linearly.

Notes for the Rp-dominant loss model plot (bottom):
There is even more severe >40dB tilt with this loss model in the first circuit. (The version with Rp on the amplifier side has a nearly flat peak response, but as I mentioned before such placement or Rp is not likely to properly reflect radiation and core losses of the inductor. Also note super-linear growth of Q with frequency in this model).

Notes for the Rs+Rp mixed loss model plot (middle):
The tilt is backward with respect to class 1 and this topology in principle can be used in hybrid feedback with class 1 oscillators but the tilt is much more pronounced than in a Vackar oscillator. Class 2 is definitely a better complimentary match for class 1 in terms of the tilt magnitude.

Hybrid feedback oscillators.

Finally, the next class represents combined (hybrid) feedback oscillators. Four oscillators were included in this study: "extended range" Vackar (fig6 in his original paper), Armstrong-Vackar hybrid, Seiler-Vackar hybrid and Hartley-Vackar hybrid.

The first circuit - "extended range" Vackar was simply redrawn on the test bench circuit diagram to highlight the pi-network feedback style, it is indeed a full equivalent to the fig.6 from his paper [1]:

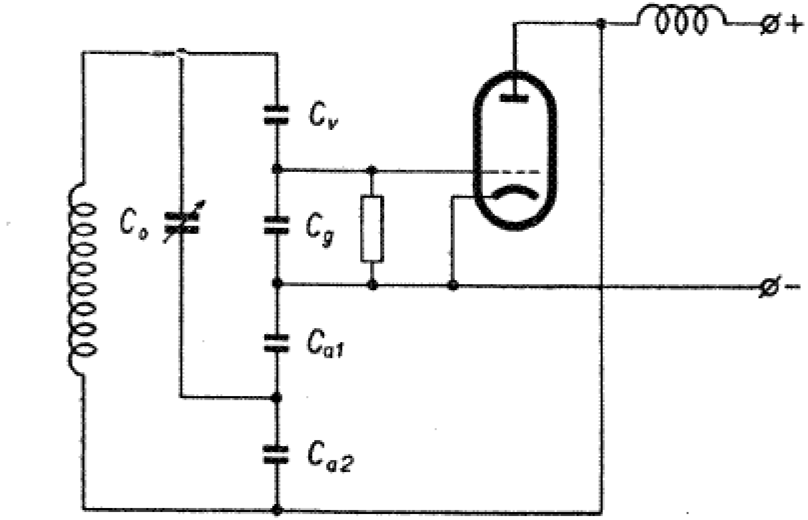
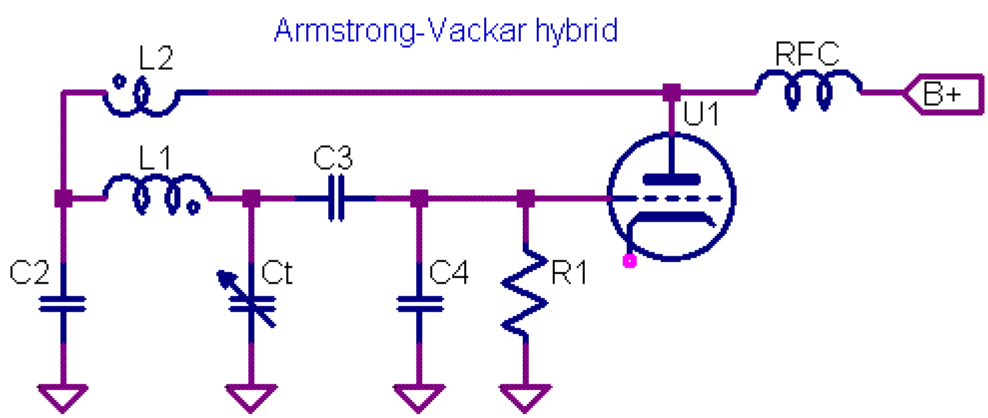


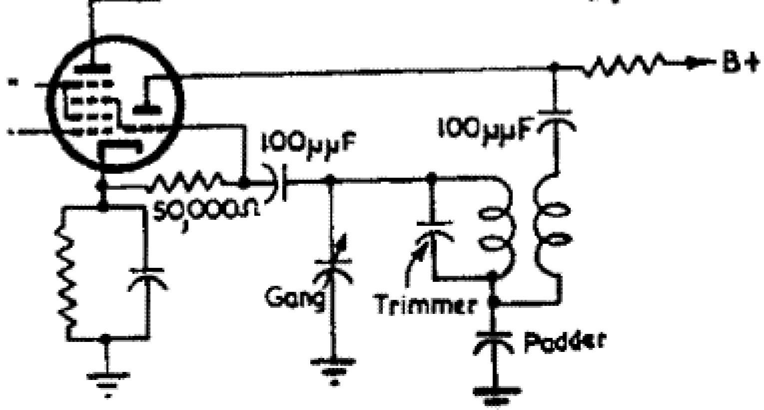
Fig. 6

I have came up with the second circuit, the Armstrong-Vackar hybrid, independently, working from the ideas expressed in the Vackar paper. Later I found a similar arrangement in the two references (see the next page).



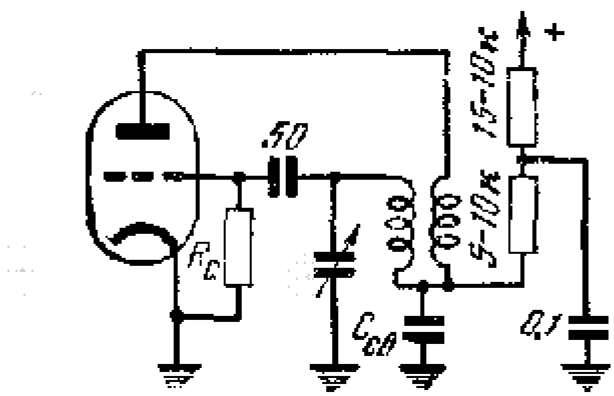
The first reference is the "padder feedback" oscillator-converter covered in RDH4 [2] (pg.1247):

FIG. 35.7



In the suggested application the circuit has one important limitation - the padder capacitor is also used for the superhet LO-RF tracking, hence this capacitor must have a specific value. This value is usually not optimal for the feedback frequency dependence compensation.

A very similar circuit is shown in a Russian radio technician textbook [3] (pg.79). The corresponding text also implies that the feedback capacitor Ссв is used for a superhet tracking, hence the circuit has the same optimization restriction. (For unknown reason, this oscillator is recommended specifically for LF band). Below is the circuit diagram from [3]:

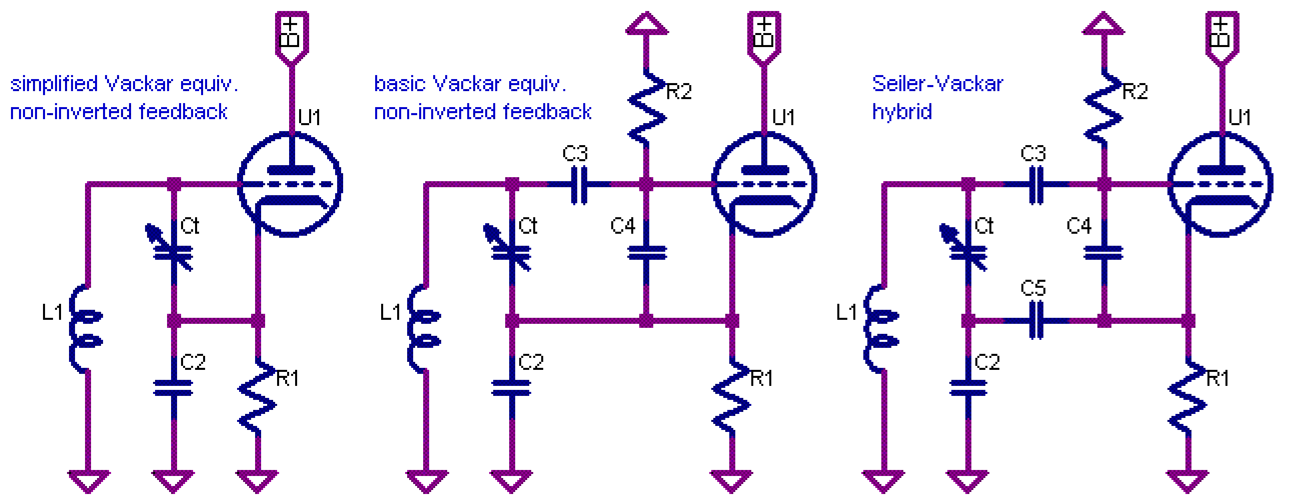


Фиг. 4-10. Гетеродин с комбинированной индуктивно-емкостной обратной связью.

If this type of circuit is used as a LO in a superhet receiver an optimal frequency compensation can be achieved by using separate feedback and padder capacitors (with a padder connected conventionally in series with a ganged tuning capacitor).

The third circuit in hybrid class test bench is a Seiler-Vackar hybrid that has some resemblance to the "extended range" Vackar but there are important differences:
1. It uses non-inverted feedback from the cathode.
2. Like in most class 1 oscillators the inductor is grounded and none of the feedback network elements are exposed to B+ (compared to Vackar circuit).
3. In both "extended range" Vackar and Seiler-Vackar the tuning capacitor Ct is not grounded. However in the proposed Seiler-Vackar circuit the Ct ground terminal is connected to the signal with much lower RF impedance making this hybrid circuit more practical - C2 on the diagram below has much lager value than the recommended value for Ca2 in the "extedned range" Vackar (fig.6).

To illustrate the origins of the Seiler-Vackar circuit consider these variations on both the simplified and basic Vackar circuits from class 2:

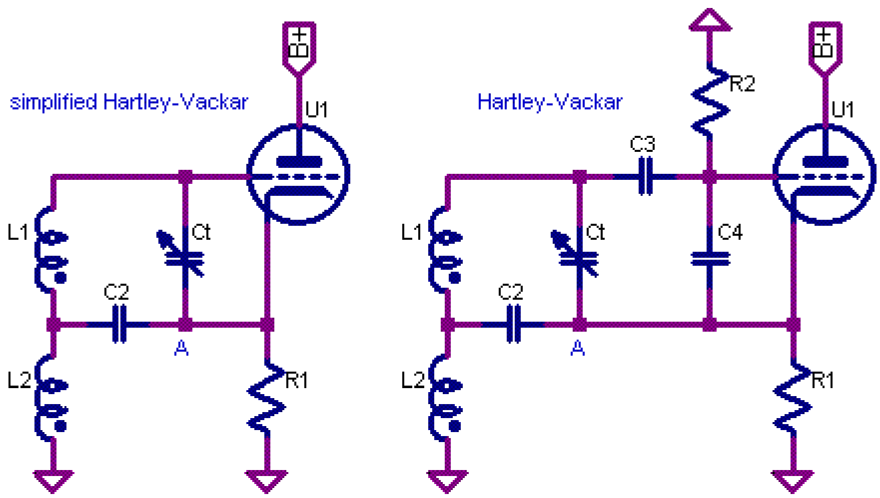


These circuits operate exactly as their Vackar counterparts from class 2 test bench diagram. They were derived by: (i) re-referencing inductor L1 to ground (instead of midpoint of Ct/C2), (ii) using non-inverting amplifier output while simultaneously applying feedback current to the opposite side of C2 (Ca in Vackar paper notation), hence maintaining the same exact phase relationship as in the original Vackar. For a fixed frequency case the first circuit is a plain Colpitts, but the placement of the tuning capacitor makes it behave exactly like a simplified Vackar in terms of the threshold behavior. The second circuit adds Vackar divider C3/C4. The third circuit adds an additional Seiler-like feedback by current injection into C5. The last circuit is included in the hybrid class test bench.

Note that all networks in this paper are completely reversible although their input and output impedance characteristics are substantially different. The former means that with an ideal VCCS amplifier there are twice as many variations possible. For example in a simplified Vackar (1st circuit on the diagram above) one can swap Ct and C2 and it will not change the frequency domain threshold behavior (this may not be trivial but it is a mere input/output swap of a Vackar pi-network). However when matched with a real amplifier that has finite and different input and output impedances these duals will behave quite differently and selecting the right one requires in depth engineering analysis.

Finally the Hartley-Vackar hybrid feedback oscillator uses the tapped inductor arrangement. It can be thought of as an extension of the first circuit on the previous page - "simplified Vackar equivalent with non-inverting feedback" but with C2 connected to the tap of L1. The hybrid feedback current flows via serially connected Vackar feedback capacitor C2 and the feedback coil section L2. Again both terminals of the tuning capacitor carry RF signal, but the node A has relatively low impedance for RF, and if connected to the tuning capacitor case, the parasitic capacitance effects can be manageable.

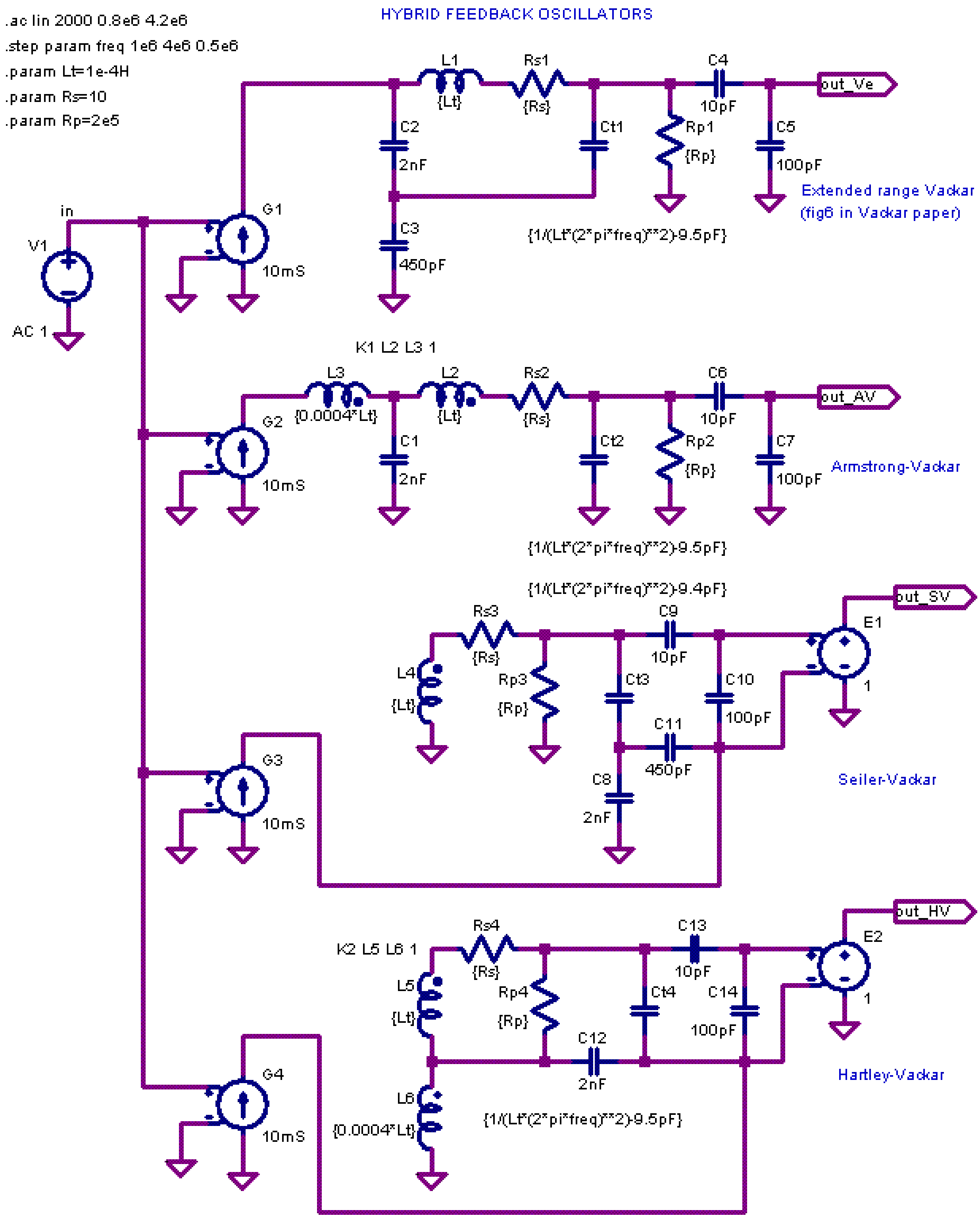
The diagrams of both simplified (no divider) and basic Hartley-Vackar oscillator are shown below:

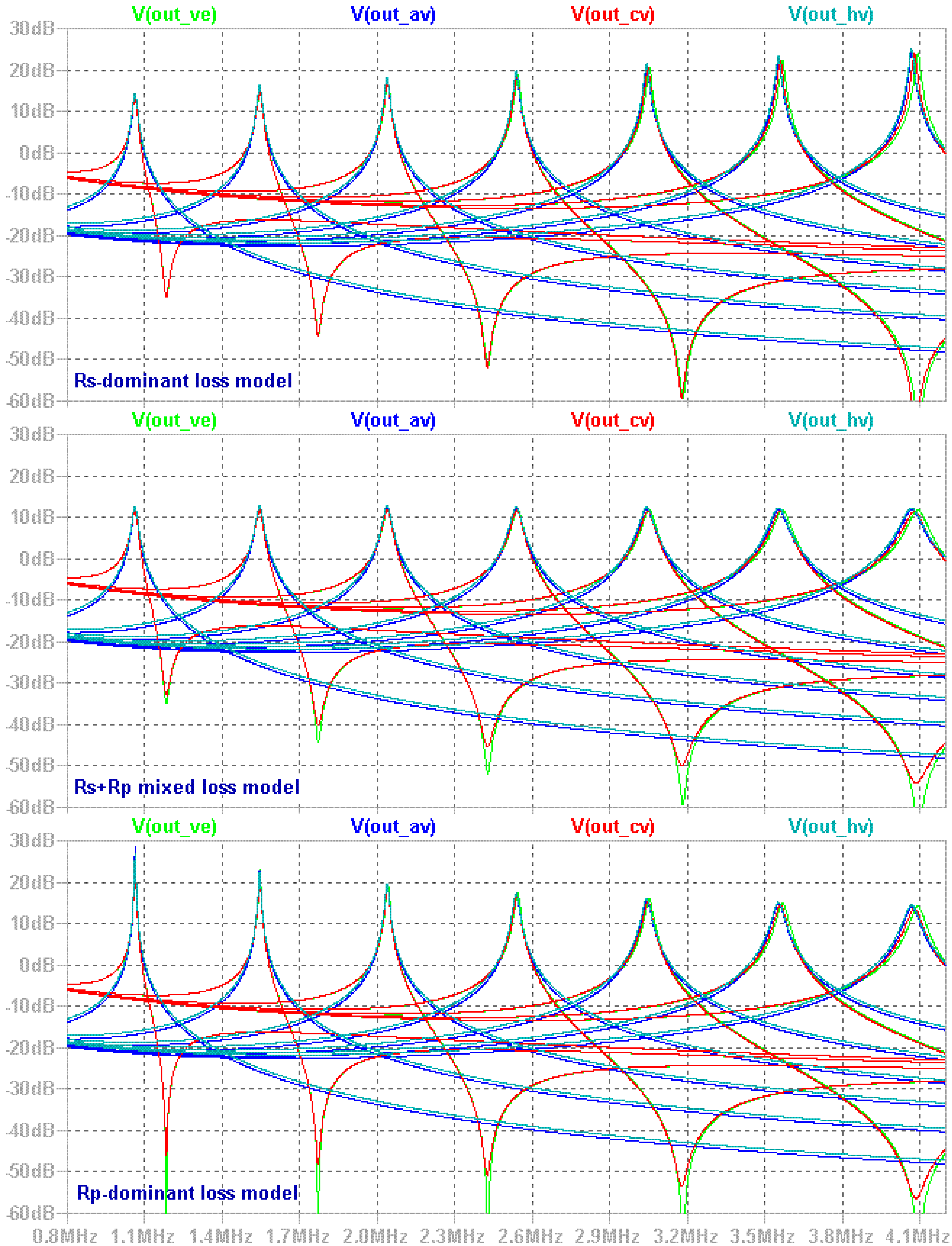


There are many other hybrid combinations possible. Some of the combinations I simulated use both inverted and non-inverted amplifier outputs for the feedback. But out of many options these four circuits were selected as being reasonably simple and representative to illustrate the hybrid feedback benefits.

The feedback network values were specifically adjusted in the simulator for nearly flat threshold with the same Rs+Rp loss model as was used for other classes. With the optimized values all hybrid class oscillators exhibit opposite tilt in the Rp-dominant and Rs-dominant loss models.

The circuit test bench and the simulation results for all hybrid class oscillators are shown on the following two pages.





V(out_ve)
V(out_av)
V(out_cv)
V(out_hv)
30dB
20dB
10dB
0dB
-10dB
-20dB
-30dB
-40dB
-50dB
-60dB
Rs-dominant loss model
V(out_ve)
V(out_av)
V(out_cv)
V(out_hv)
30dB
20dB
10dB
0dB
-10dB
-20dB
-30dB
-40dB
-50dB
-60dB
Rs+Rp mixed loss model
V(out_ve)
V(out_av)
V(out_cv)
V(out_hv)
30dB
20dB
10dB
0dB
-10dB
-20dB
-30dB
-40dB
-50dB
-60dB
Rp-dominant loss model
0.8MHz
1.1MHz
1.4MHz
1.7MHz
2.0MHz
2.3MHz
2.6MHz
2.9MHz
3.2MHz
3.5MHz
3.8MHz
4.1MHz

V(out_ve) represents the frequency response magnitude for the "extended range" Vackar oscillator;
V(out_av) - for Armstrong-Vackar hybrid feedback oscillator;
V(out_cv) - for Seiler-Vackar hybrid feedback oscillator;
V(out_hv) - for Hartley-Vackar hybrid feedback oscillator;

Notes for the Rs-dominant loss model plot (top):
Slight forward tilt as expected.

Notes for the Rp-dominant loss model plot (bottom):
Slight backward tilt, again as expected.

Notes for the Rs+Rp mixed loss model plot (middle):
Nearly perfect zero-tilt for a mixed Rs+Rp model. The LC network values can be tuned to achieve exactly zero tilt for any given combination of Rs and Rp (not possible for classes 1, 2 and 3 alone).
Note nearly constant Q for the mixed load model.

Verification in the hardware.

The simplest practical circuit to verify the tilt compensation in any oscillator topology is a 100-year old regenerative receiver circuit - it is essentially an oscillator with a variable gain amplifier. The onset of an oscillation occurs when regeneration (amplifier gain) is set such that $Q(f)*P(f) >= 1$. The experiment does not require an antenna, one can simply observe the onset of an oscillation with a scope while varying amplifier gain and frequency. (If the circuit is used as a true receiver then RF buffer is highly desirable since the antenna has substantial effects on the resonant circuit losses and may change the tilt behavior significantly.)

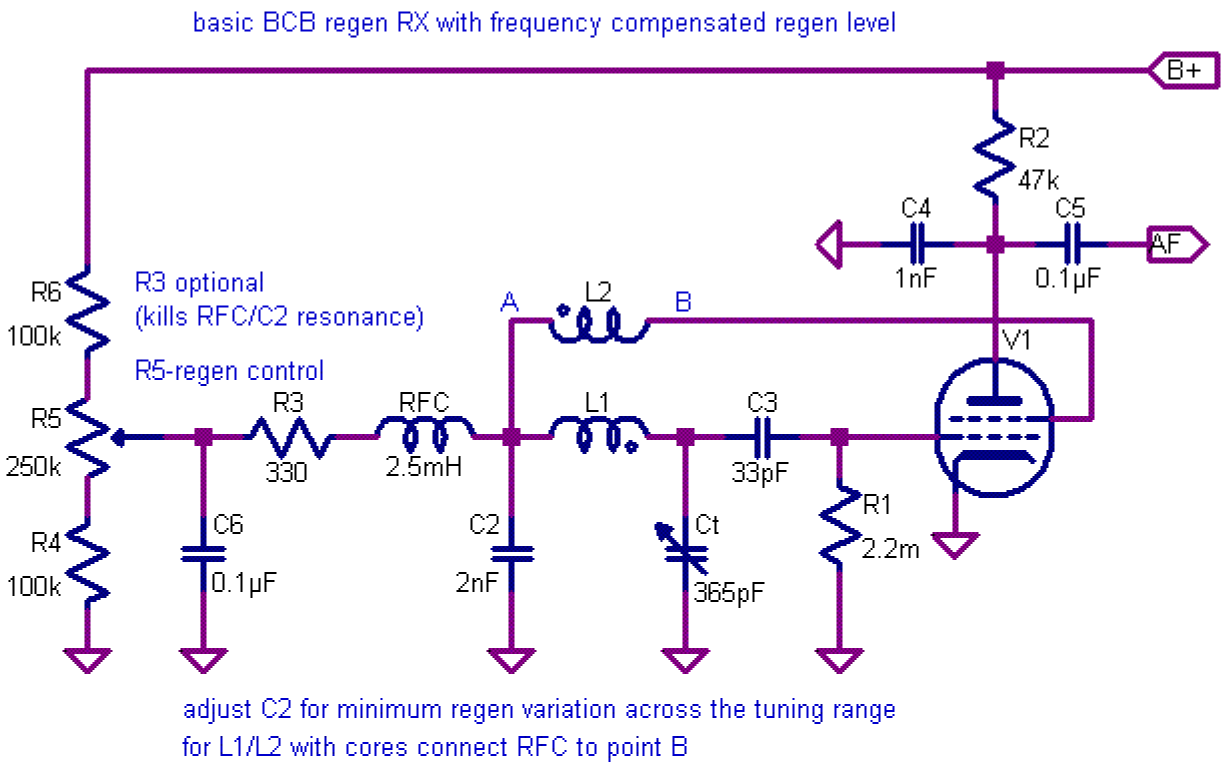
Practical experience shows that most conventional regenerative receivers require substantial adjustment of the regeneration level as you tune across the band. For example most common regens using class 1 oscillators - Armstrong / Hartley / Colpitts topology usually require advancing the gain (regeneration) at the low side of the band. On the other hand in the circuits based on the class 2 and 3 oscillators the threshold has the opposite tilt and the gain must be advanced at the high end of the tuning range. This closely matches the simulation results with a mixed loss model Rs+Rp.

All three hybrid feedback oscillator prototypes built in this study confirm that after properly balancing the two feedback paths the oscillation onset happens at exactly the same gain settings at the two opposite frequencies of the tuning range (or any two frequencies of choice) and that the remaining non-linear term of the threshold dependency is quite small.

Due to: (i) my interest in the history of the vacuum tube circuitry, and (ii) the fact that Vackar came up with the first tilt compensated oscillator circuit in the vacuum tube era, I decided to verify the idea using vacuum tubes. Nevertheless I want to emphasize that the concept itself is equally applicable to both solid state design and to modern applications such as VCO in PLL circuits.

Test prototype 1.

The first prototype was based on a battery-powered dual 1T4 tube MW regenerative receiver (I've built it with kids some time ago). The board was easy enough to modify so I used it as a testbead. Armstrong-Vackar hybrid feedback circuit is particularly easy to implement - any standard Armstrong regen can be modified to operate with a combined feedback. The circuit is slightly unconventional in that it uses screen grid as a virtual anode for the RF feedback. This is mostly for the legacy reasons since the donor RX was done this way, but it does add some convenience of separating the RF and AF paths, which is helpful when the resonant tank is not grounded for AF.



C2 value should be be adjusted as follows:
Starting value for C2 should be around 10x Ct max. Starting number of turns in L2 - slightly less than in a conventional Armstrong regen (if unsure of the number of turns in L2 - try Armstrong-only regen first). Once you get the amount of total feedback approximately right then tweak C2 for zero tilt. Decrease the value of C2 if the regeneration level potentiometer R5 needs to be advanced more at the lowest frequency. Increase the value of C2 if the opposite is true. There is a clear optimum value for C2 that corresponds to the exactly zero tilt.

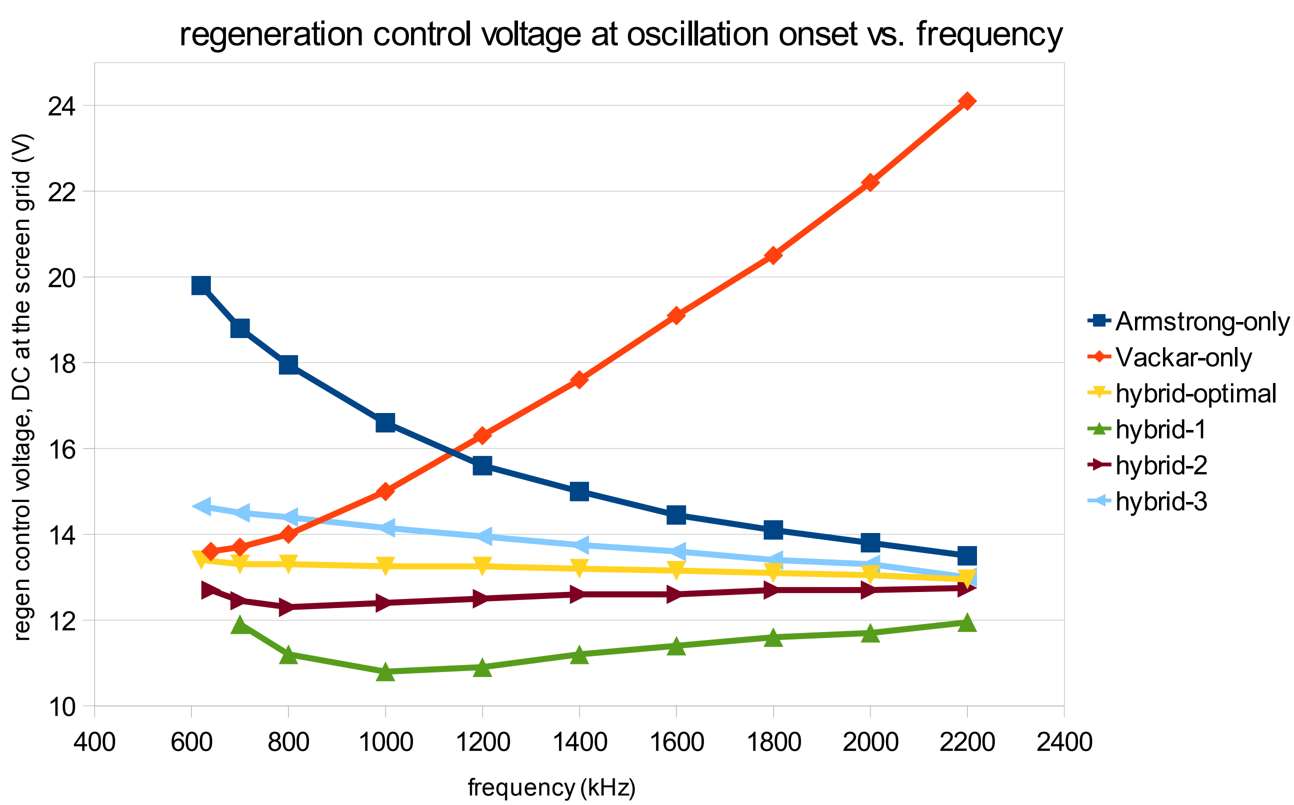
Via C2 adjustment the tilt was indeed zeroed out. This receiver did have some uncompensated higher order terms in the oscillation threshold frequency dependence and required minor decrease in the amount of regeneration in the middle of the BCB band, but the amount of remaining adjustment was much lower than in a conventional class 1 oscillator based regen. The shortwave prototype that will be discussed later had even smaller higher order terms. It may be related to the fact that this MW receiver has a fairly large air coil with measurable radiation and resistive losses, while SW version uses a high-

Q toroidal inductor.

The next diagram shows actual (measured) dependence of the regeneration control voltage at the oscillation threshold (DC at grid2) vs. frequency for several configurations of the circuit above:
1. Armstrong-only feedback (C2=0.1uF);
2. Vackar-only feedback (C2=1000pF, L2 bypassed);
3. Optimal hybrid feedback, minimum tilt (C2=1470pF);
4. Sub-optimal hybrid feedback case 1 (C2=470pF);
5. Sub-optimal hybrid feedback case 2 (C2=1000pF);
6. Sub-optimal hybrid feedback case 3 (C2=2470pF);

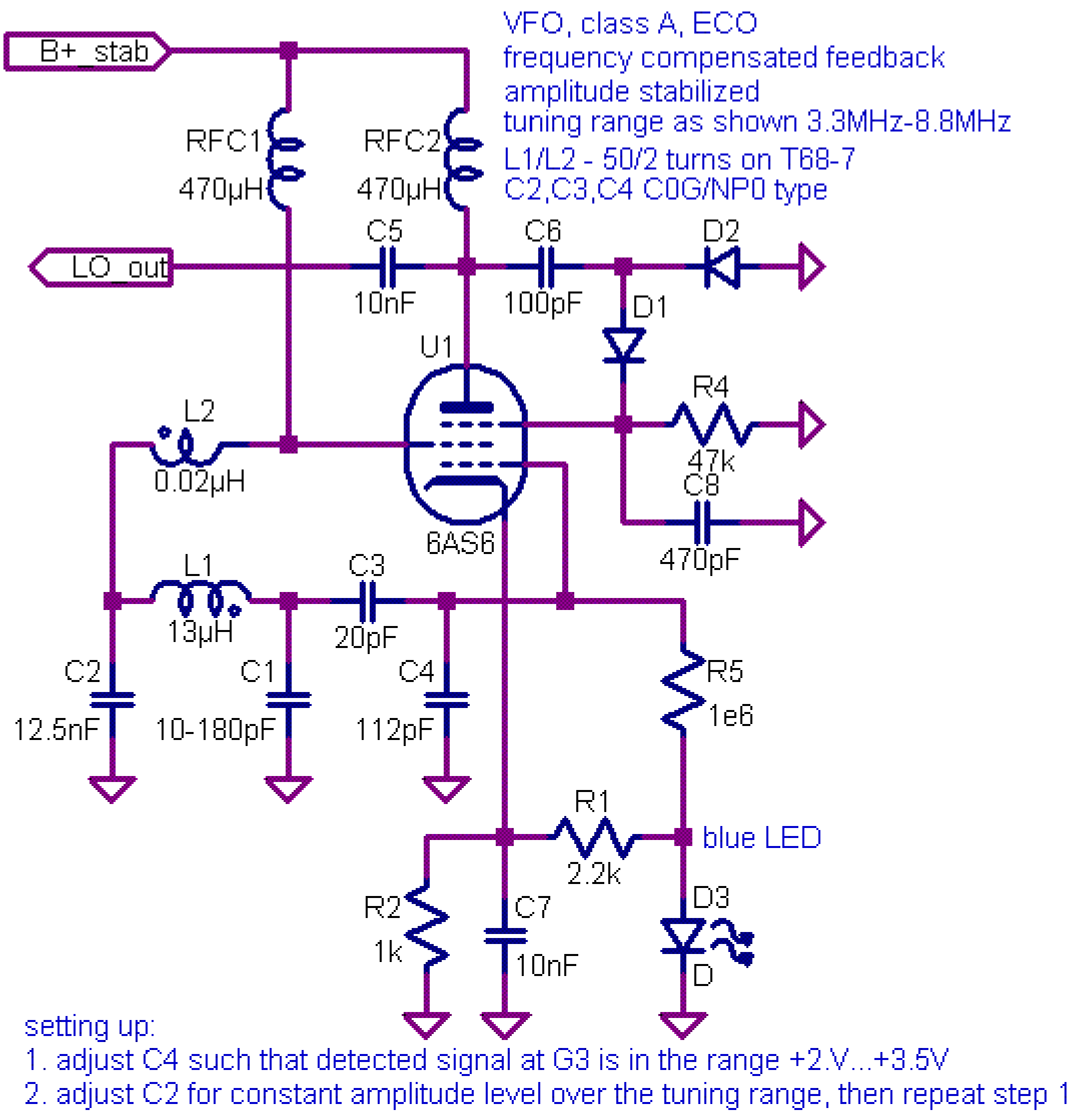
Note that direct comparison with the simulation results is not possible for two reasons:
1. Amplifier gain is the inverse of the LC network frequency response.
2. The regeneration control DC voltage has somewhat nonlinear relationship with the amplifier gain. However the relationship is monotonic and smooth, so the character of the dependence is preserved.

It can be observed that:
1. Both Armstrong-only and Vackar-only feedback configurations exhibit noticeable tilt in opposite directions, fully consistent with the simulation results.
2. There is an optimal value of C2 (around 1500pF) that minimizes the tilt.
3. The remaining uncompensated non-linear terms are small with the optimal value of C2.
4. The tilt sensitivity to C2 adjustment is not sharp. (SW prototype has more sensitive tilt adjustment)
5. There is some dependence of the non-linear terms on the value of C2. (SW prototype exhibits less residual threshold non-linearity around the optimal C2 value).



Test prototype 2.

Success of the tilt compensation in the simple prototype described above encouraged me to work on a second prototype. This time I built a single tube, class A, amplitude stabilized, electron coupled oscillator using a single 6AS6 dual control pentode. The circuit diagram is shown below:

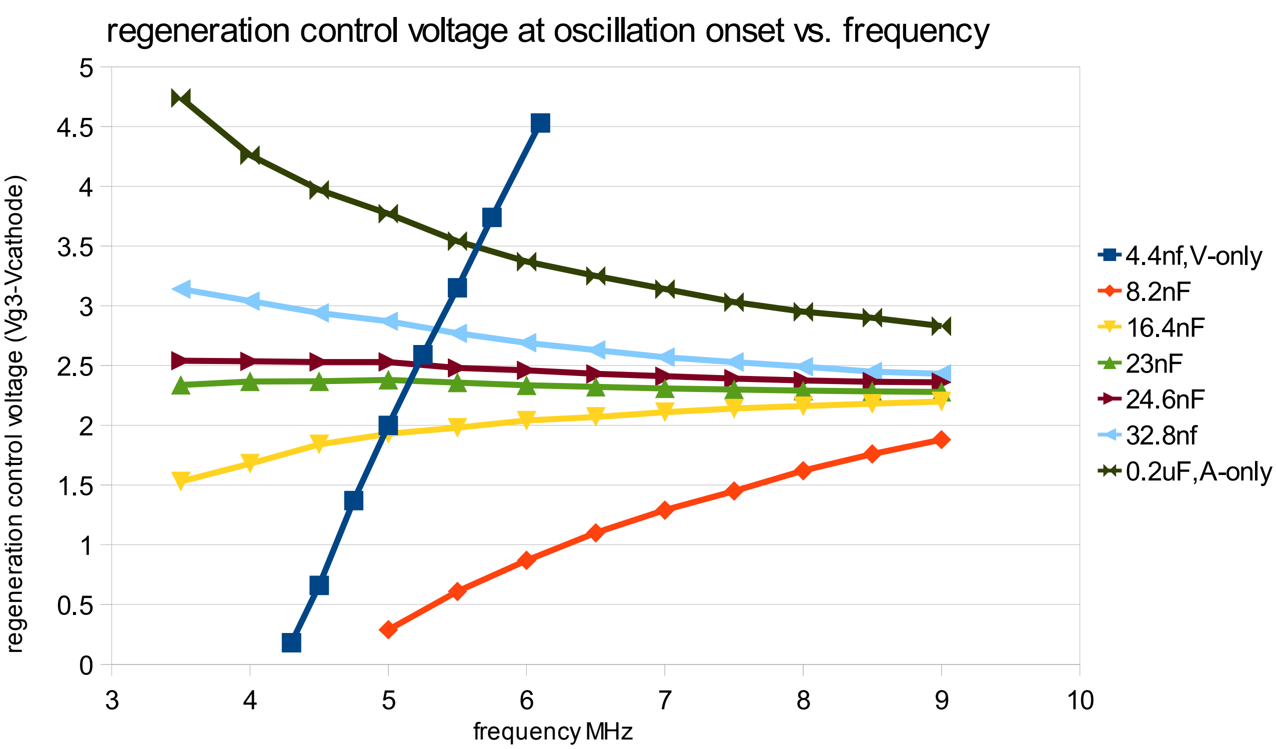


The oscillator gain is controlled by applying the rectified RF voltage to G3. Note that the oscillator feedback is taken from the screen grid, hence raising the G3 voltage reduces the feedback gain (this is specific to the dual control pentodes like 6AS6). The LED network was used to raise the cathode potential to allow larger initial negative bias for G3, hence providing higher oscillator start-up gain. The method of the amplitude control via G3 has two benefits: (i) it offers more linear RF amplification compared to biasing G1, and (ii) it does not load the tank circuit directly with an AGC detector. The circuit has extremely low tank loading hence should have very low phase noise (although I have no tools to measure and verify it directly).

After carefully adjusting the values of L2, C2 and C4 a nearly flat oscillation threshold from 3.3MHz to 8.8MHz (2.65x frequency span) was obtained. This has been verified across the tuning range by:
1. measuring detected voltage at G3 while tuning the oscillator across the band;
2. adding the manual gain control (by adjusting the detector reference voltage) and monitoring the gain (G3 voltage) at which the oscillation starts while tuning the oscillator across the band.

The actual measurements for the graph below were taken from the prototype SW receiver discussed in the next section, reverting it back to the VFO arrangement by removing the RF buffer vacuum tube. As opposed to the BCB regen circuit, increasing the DC potential at G3 reduces the feedback gain, therefore to make the graphs comparable to the BCB regen graphs, all Y-axis values were flipped by computing Vcath-Vgrid3 series, where Vcath is nearly constant at 4.95V. Note that the amplifier gain is the inverse of the LC network frequency response shown in the simulation results.

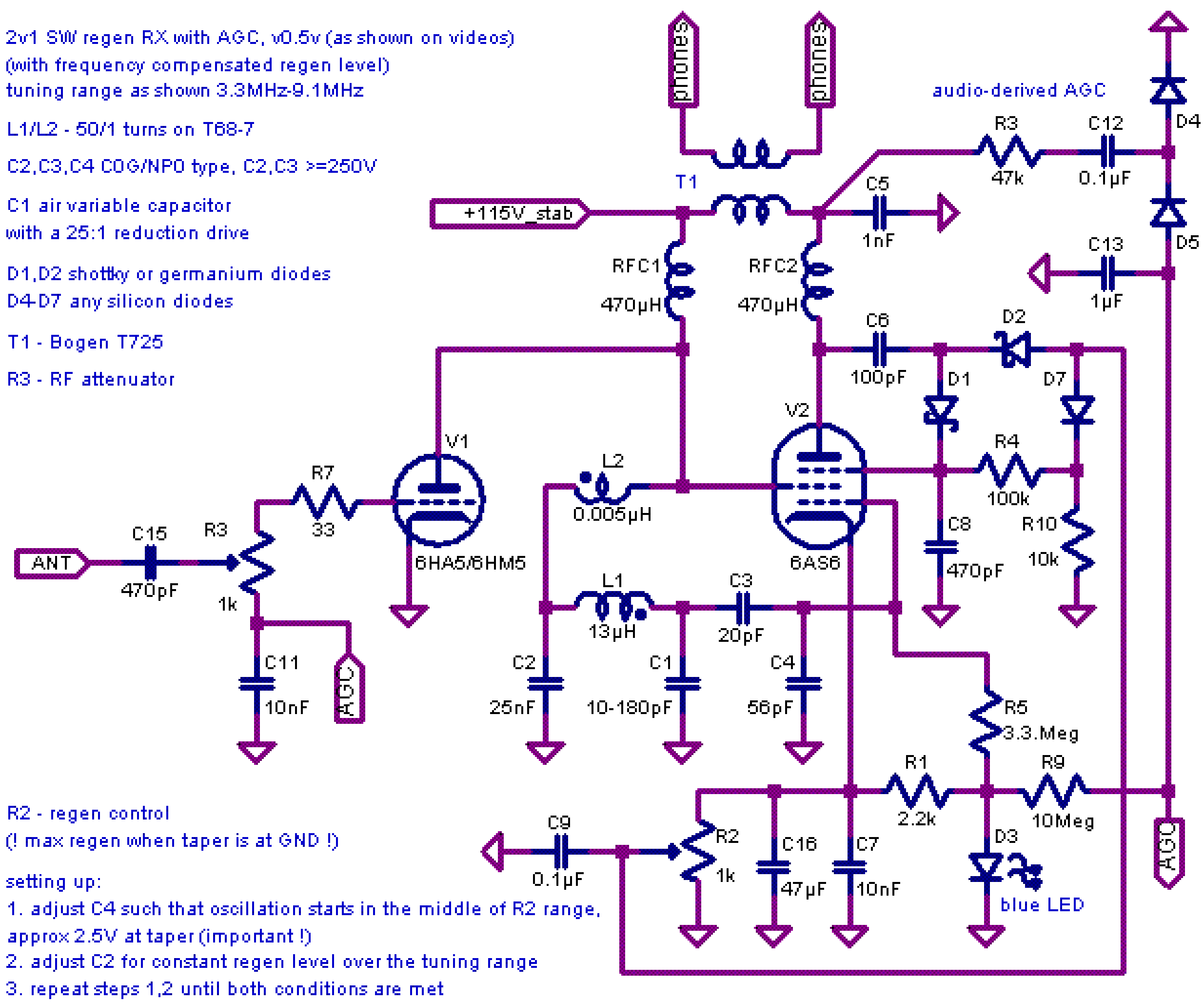
The graphs illustrate the following feedback configurations:
1. Vackar-only feedback (C2=4.4nF, L2 bypassed). There was not enough gain control voltage range for testing the entire tuning span, so I had to limit the experiment to 4.3MHz-6.1MHz.
2. Armstrong-Vackar hybrid (C2=8.2nF)
3. Armstrong-Vackar hybrid (C2=16.4nF)
4. Armstrong-Vackar hybrid (C2=23nF)
5. Armstrong-Vackar hybrid (C2=24.6nF)
6. Armstrong-Vackar hybrid (C2=32.8nF)
7. Armstrong-only feedback (C2=0.2uF)



While the full tilt compensation was expected from this circuit, a pleasant surprise was that the higher order terms of the threshold frequency dependence were very small (noticeably smaller than in the BCB circuit). For a practical regen/VFO/VCO (except maybe some test and measurement applications) they are negligible.

Test prototype 3.

The VFO circuit above was already instrumented with the manual gain control for the threshold study so it was a natural progression to try it as a shortwave regenerative receiver. The first thing I immediately noticed that connecting antenna to the resonant circuit severely affected the tilt compensation. This was expected due to both capacitive loading of a short random wire antenna and greatly increased radiation losses in the resonant circuit. To circumvent this I added an isolation RF amplifier between the antenna and the VFO/detector. Additionally the output coupling of both detector and RF amplifier to the resonant circuit was further reduced (compared to the VFO prototype above) and the resulting loss of gain compensated by increasing the detector input coupling (by decreasing C4). The resulting circuit had a tuning range span from 3.3MHz to 9.1MHz (about 2.75x). The circuit diagram is shown below:



In this circuit the total sum of: (i) detected AF voltage, (ii) RF carrier magnitude, (iii) RF self-oscillation magnitude, and (iv) regeneration control voltage is applied to G3. AF voltage gets some reflex amplification before reaching the T1 primary. Combined RF magnitude of the carrier and the self-oscillation control the loop gain of the detector/oscillator. I am skipping other circuit details and

testing/debugging issues that are not related to the dual feedback network. For those interested a full detailed account of this circuit creation history can be found at this forum:
http://www.theradioboard.com/rb/viewtopic.php?t=3714

The end result was a surprisingly well behaved regenerative receiver. It has a very good frequency stability and does not require adjusting regeneration level as you tune from 3.3MHz to 9.1MHz which is highly unusual for a conventional regen. The oscillation threshold level is so flat that it is completely feasible to have just two fixed regeneration presets - one for AM and a higher one for the SSB/CW modulation types. To illustrate these somewhat unconventional results I put several videos on youtube. The two most relevant to the subject of this discussion are:
http://youtu.be/Frg5RYz84gs
showing the scope measurements of the oscillation onset while tuning the circuit, and
http://youtu.be/zKRpVqzFGYE
showing the receiver operation by setting the regeneration level just below the oscillation threshold and tuning the receiver across the entire range (2.75x frequency span) without touching the regeneration control.

Conclusions and further study.

Conclusions:

1. The simulation and the prototypes confirm that it is possible to build a capacitively tuned oscillator that has full compensation of the linear term of the oscillation threshold frequency dependance over the wide tuning range using lossless methods alone (no additional resistive tank damping beyond intrinsic losses). Practical limitations on capacitively tuned oscillators limited the prototypes to slightly below 3x tuning range.
2. The tilt linear term compensation can be achieved for any combination of intrinsic linear loss parameters Rs and Rp. The final optimization can be done in place.
3. Despite the complex non-linear dependence of the loss parameters Rs(f) and Rp(f) on frequency the higher order terms of the threshold frequency dependence after the tilt elimination are very small.
4. There exists a large class of combined (hybrid) feedback oscillators having these properties. Only the four examples were simulated but many more are possible.
5. The method of lossless tilt compensation is suitable not only for "antique" circuits but also for the modern solid state design and applications.

Possible further study directions:

1. Compensating tilt in the inductively tuned oscillators.
2. Testing solid state circuits using the hybrid feedback topology: regen/VFO/VCO/PLL
3. Trying to scale the idea to the UHF/microwave frequencies and the transmission line oscillator design.
4. Perform an accurate phase noise study of the hybrid feedback oscillators.
5. Expanding the topology search: other hybrids, split (inverted+non-inverted) feedback, etc.
6. Simplifying the in-circuit tilt adjustment - possibly using two separate amplifiers for the two feedback branches and a potentiometer gain balancing between the branches.

References:

[1] http://n1ekv.org/Oscillators/Vackar_wholepaper.pdf
[2] Radiotron Designer's Handbook. Fourth Edition, 1953 (a.k.a. RDH4)
[3] В.К. Лабутин, "Книга Радиомастера", издание 1955 (in Russian)

Acknowledgements:

I would like to express thanks to N1EKV [? Byron Blanchard ?] for his effort making the original Vackar paper available to a wide engineering and hobbyist audience. The paper initiated my interest in the frequency compensated feedback.

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