Design of a Regenerative Receiver for the Short-Wave Bands
A Tutorial and Design Guide for Experimental Work

Part II

Ramón Vargas Patrón
rvargas@inictel-uni.edu.pe
INICTEL-UNI

Summary

Part I of this study addressed a comprehensive, but not exhaustive, presentation of the Q factor of a tuned circuit. Expressions for the quality factor of lossy reactances and parallel resonant networks were obtained based upon the stored and dissipated RF energy concept. Its relationship with selectivity, -3dB bandwidth and resonant frequency was also discussed.

Part II will briefly comment topics related to:
-power supply regulation
-heating problems producing operating-point drift
-temperature dependence of active-device parameters and influence on circuit gain
-low DC power dissipation in the active device for thermal stability
-low RF-loss materials for coil forms and front panels
-ground planes
-mixer-type detection of SSB signals in the active detector stage
-frequency stability requirements for acceptable SSB detection
-coupling the antenna-ground system to the receiver and sensitivity-selectivity trade-off
-BJT and JFET RF types suitable for smooth gain control in regenerative receivers.

In the present report, a simple Armstrong-type regenerative receiver for the short-wave bands capable of tuning from 3MHz to 12MHz will be presented as a design exercise. After field-testing a prototype of this receiver, the author considered it convenient to conduct a comprehensive theoretical study of the circuit for interested readers, the same that has been included in this paper. The Armstrong receiver topology was chosen because of its straightforward design procedure, and high rate of success reported by radio experimenters worldwide.

Comments on Device Behavior and Materials Prior to Selecting a SW Regenerative Receiver Topology

Regenerative receivers can demodulate DSBC AM signals (double side band with carrier present) as well as SSB (single side band) signals, where the carrier and one side band have been suppressed. These receivers may use a detector-amplifier, where no external demodulator diode is needed. For this purpose, the operating point of the RF-amplifying device is positioned near the bottom of the transconductance characteristic curve. In the case of DSBC, this gives rise to asymmetrical amplification of the AM signal and a DC output current that varies following the envelope of the AM modulated carrier. The varying current produces a voltage drop across a load resistor from which finally the modulating signal is extracted with low distortion.

Regulated DC power supplies are recommended for working out SSB signals. These are demodulated (detected) after a non-linear mixing process occurring in the active device. In this mode of operation, the detector-amplifier is put into oscillation, the frequency depending on external timing components and transistor parameters. The latter are operating-point dependent and ultimately, a function of the DC power supply. This is most certain for BJTs (Bipolar Junction Transistors), where collector currents exhibit high sensitivities to power supply variations.

In the case of JFETs (Junction Field-Effect Transistors), quiescent-point drain currents depend strongly upon V_{GS} (gate-source DC voltage = I_{SOURCE} x R_{SOURCE}) and little upon V_{DS} (drain-source DC voltage), if the latter exceeds the V_{GS} pinch-off voltage. So, for JFETs, less stringent requirements for power-supply regulation are needed.

Transistor parameters are also temperature dependent. Temperature elevation in a device produces some operating-point drift, situation being more noticeable in a BJT. This could be due to poor ventilation or excessive power dissipation in the device. This explains one reason for selecting a low quiescent-point drain or collector current in a regenerative receiver. In the non-oscillating mode an operating-point drift will show up as a change in the volume of a received program, distortion, or as circuit detuning. In the oscillating mode the problem will show up as a frequency drift of the oscillation signal.

On the other hand, regeneration produces *Q* magnification and a shift of the resonant frequency from the lossy situation value to that of the ideal lossless case. So, if regeneration level is adjusted, some detuning is to be expected. In this sense, it is advisable to use coils of the highest possible *Q* for least frequency shift. Likewise, construction methods for high-Q coils should not be overlooked. Low RF-loss coil forms of common use by radio experimenters for HF frequencies include materials as polystyrene pipe and plastic 35-mm film containers for analog cameras. The film containers are found to be manufactured from HDPE (High Density Polyethylene). Front panels for non-critical radio circuits and small boxes can be constructed from clear acrylic sheets. They are reasonable low-loss materials.

Short-wave coils can be made close-wound or non-close wound using enamelled solid copper wire or PVC-insulated solid hook-up wire, the number of turns, wire gauge and coil diameter obtained from formulas detailed in technical books, or using software applications found on the web. A nice design aid written by Dan Petersen can be found at: http://crystalradio.net/professorcoyle/

Another important reason for selecting a low operating-point collector or drain current is the device's g_m (transconductance) tendency to increase with output current, making it very difficult to obtain a smooth regeneration control. Small g_m's facilitate gain control. Hence, lower operating-point currents. The idea is to have perhaps some 60% of the required gain for oscillation with the correct collector/drain current already established, the extra 40% obtained through the use of positive feedback. If one already has 95% of the needed gain, obtaining the extra 5% will require VERY little adjustment of the regeneration control, and the whole thing may start whistling or giving out unpleasant squealing sounds with the slightest adjustment. This can be quite annoying to the operator if he is only trying to get a little more amplification for a weak radio signal.

We must emphasize the necessity for having a good low-impedance ground plane for the return of RF and AF currents and for keeping AC fields confined in the components area and avoid line forces crossing the operator’s body or nearby objects underneath the circuit layout or on the sides. A good ground plane will help to make of a receiver a piece of gear very easy to operate, free of hand-capacitance effects.

Low-impedance ground planes can be implemented using techniques that include the time-proven Manhattan construction style and the ugly-bug style, for soldered and complex projects, and the protoboard or solderless breadboard prototyping method with an aluminum sheet underneath as a ground plane, for the insertion-wire method. To be effective, the protoboard’s ground bus should be solidly connected to the ground plane with a solder lug, screw and respective washer and nut. Variable capacitors, potentiometers, switches, selectors and headphone/loudspeaker jacks should be mounted on an aluminum panel making intimate mechanical and electrical contact with the aluminum ground plane. It must be born in mind that solderless breadboards exhibit some 3pF of stray capacitance between adjacent connecting rows. The author has been able to successfully work regenerative receivers on protoboards up to 28MHz receiving broadcast and amateur SW radio signals.

The above paragraphs apply to all receiver topologies. Experience shows that operating-point collector/drain currents of 50uA......200uA are enough for a reliable operation of the detector-amplifier stage, no matter the selected topology. Ultimately, the smaller the operating-point current is, the quieter the amplifier will be (regarding circuit noise). As an example, in the particular case of an Armstrong-type receiver, the available g_m, the in-loop RF transformer ratios and the value of the feedback capacitor employed for regeneration control dictate the magnitude of the needed current. Also, the feedback capacitor, which is ultimately the manual RF gain control, should be maintained somewhere between 20% and 50% of its maximum value throughout the whole tuned band for maximum signal amplification. It's not comfortable to have to change the capacitor's setting extensively when tuning from minimum to maximum frequency.

Here proves an effective aid the use of the Vackar's approach to reduce the rotation-angle of the variable capacitor between ends of the band.

In the Reference section the reader will be able to find useful literature regarding the excellent contributions to radio transmitter and receiver technology made by Engineer Jiri Vackar from the Czech Republic.

Mixer-Type Demodulation of SSB Signals in the Detector-Amplifier Stage of a Regenerative Receiver

The detection of SSB signals requires the receiver working in the oscillating mode and generating a signal of constant amplitude and frequency, this is, acting as a BFO (Beat Frequency Oscillator). Conditions for the latter will be met as long as the oscillation conserves its integrity, i.e., its amplitude and frequency. In other words, it must not be related in any way to the SSB signal impressed on the tank circuit by a passing radio wave. No frequency-pulling effect should occur, and in this sense, strong oscillations will be needed so the receiver doesn´t frequency-lock to the incoming signal.

SSB signals available at the input of the receiver have a large dynamic range. It can be a local nearby amateur signal or a transcontinental link. It is advisable then to have some type of variable attenuator in series with the antenna lead-in wire in order to reduce input-signal strength levels to the point where frequency-pulling effects disappear. Signals with insufficient attenuation give rise to the characteristic “quack-quack”- like sounds in the retrieved audio. A 150-pF air-dielectric variable capacitor is a good choice for an attenuator and should be mounted isolated from the front metal panel, or signal loss may occur.

Let $f(t) = A\cos(\omega_c + \omega_m)t$ be the SSB signal and $s(t) = B\cos(\omega_c + \Delta\omega)t$ the receiver’s oscillation, both co-existing in the tank circuit. Detection is of the square-law type, after a mixing process carried out by the nonlinearities of the active device:

$$p(t) = [f(t) + s(t)]^2$$

Then:

$$p(t) = \left[A\cos(\omega_c + \omega_m)t + B\cos(\omega_c + \Delta\omega)t\right]^2$$

or equivalently,

$$p(t) = \frac{A^2}{2}\left[1 + \cos 2(\omega_c + \omega_m)t\right] + \frac{B^2}{2}\left[1 + \cos 2(\omega_c + \Delta\omega)t\right]$$
$$+ AB\left[\cos(2\omega_c + \omega_m + \Delta\omega)t + \cos(\omega_m - \Delta\omega)t\right]$$

After removal of RF frequency components from the output of the detector-amplifier we are left with:

$$D(t) = AB\cos(\omega_m - \Delta\omega)t$$

We observe that the spectrum of the modulating signal has been recovered shifted down in frequency by an amount $\Delta\omega/2\pi$ Hertz. There the importance of the correct adjustment of the frequency of the local oscillation, which may be done by slightly modifying bias conditions, feedback or main tuning. Accordingly, the oscillation signal must have good frequency stability and quality components should be selected for the tuning tank and feedback path.

Although the available RF power from a solid-state oscillating regenerative receiver is usually very small, less than 1mW, it is advisable to very loosely couple the antenna to the receiver in order to avoid interfering with other amateur or private services. A second solution is to add a buffer or isolation amplifier between the antenna and the input to the receiver. Usually a common-base or common-gate amplifier is used, if required.

Reduced tank-circuit loading by the antenna-ground system is another advantage arising from a loosely-coupled antenna. Loading affects selectivity in bad ways and may produce dead spots in the received band. However, a loosely-coupled antenna may reduce the receiver’s sensitivity. A compromise should be made.

BJT and JFET RF Types Suitable for Smooth Gain Control in Regenerative Receivers.

Smooth regeneration control is an appreciated feature of a well-designed regenerative receiver. It may be identified by the slow-paced entrance of the receiver to the oscillating state while adjusting the feedback control for increased gain. The onset of oscillation can be recognized by a hushing sound coming out of the headphones or loudspeaker. In a receiver with smooth operation, the hushing sound will gradually increase in intensity as feedback is augmented. When rotating the panel knob for increased regeneration, the point where oscillation starts has to be the same where it fades out in the returning direction. There has to be no hysteresis.

As mentioned before, small values of the transconductance g_m of the active device will facilitate gain control. However, not every BJT or JFET RF type will render adequate for operation in a regenerative receiver designed for smooth control. Best adapted to this labor are those transistors having rather low collector/drain output resistances. These devices have been specially manufactured for operation in circuits where matching to 50-ohm impedances are likely to occur. The regenerative receiver to be presented in the next section has been designed around the BJT 2N3904 type. While testing several samples in the detector-amplifier stage the author found specimens that would refuse to work under the statements given in the above paragraph. Vintage 2N3904’s manufactured by Motorola were unsuccessful in this sense. However, transistors marked as 2N3904 B331 from unknown manufacturer worked satisfactorily. Tested in a DMM for H_{FE}, both groups gave figures above 200. The transistor’s DC gain, then, is not an issue at all. It’s AC behavior what makes the difference. We may conclude that for reduced sensitivity to positive feedback changes (for a smooth gain control), we need small values for the transistor's transconductance g_m and devices having lower quiescent-point output resistances (larger values for h_{oe}). Needless to say, h_{oe} should continue to increase with current along the AC operating path.

The following information will be of great interest to radio experimenters working with JFETs in RF stages:

Text below and Fig.1 are from "The Triode Emulator" thread, reply #17 on October 06, 2011 http://www.diystompboxes.com/smfforum/index.php?topic=93889.msg808690#msg808690

The basic equation that governs the behavior of a JFET in the saturated regime is so only for "(infinitely) long channel" JFETs. Some production JFETs are "long-enough channel" to follow the equation closely.

A short-channel JFET has lower output resistance, but also higher gain in low-impedance loads. Short-channel JFETs are popular in RF circuits where all loads are low-impedance and gain is essential.

We often can't know which type "our" JFET is. Back around 1980 the manufacturers published some clues in their data-sheets but the JFET market has been stagnant for years and there is little new data, and not much old data.

The text tells us that the designer has to deal with two types of RF JFETs (Junction Field-Effect Transistors), the majority of times not knowing specifically if they are "short-channel" or "long-channel". There can be sensible differences between specimens of the same type number coming from different manufacturers. Actual operation in a circuit will tell if the selection of a particular device was correct. Short-channel JFETs are best choices for smooth gain control. Two types have been tested by the author, the MPF102 and the J310. They worked well in SW regenerative receivers. Both are N-channel JFETs for VHF work.

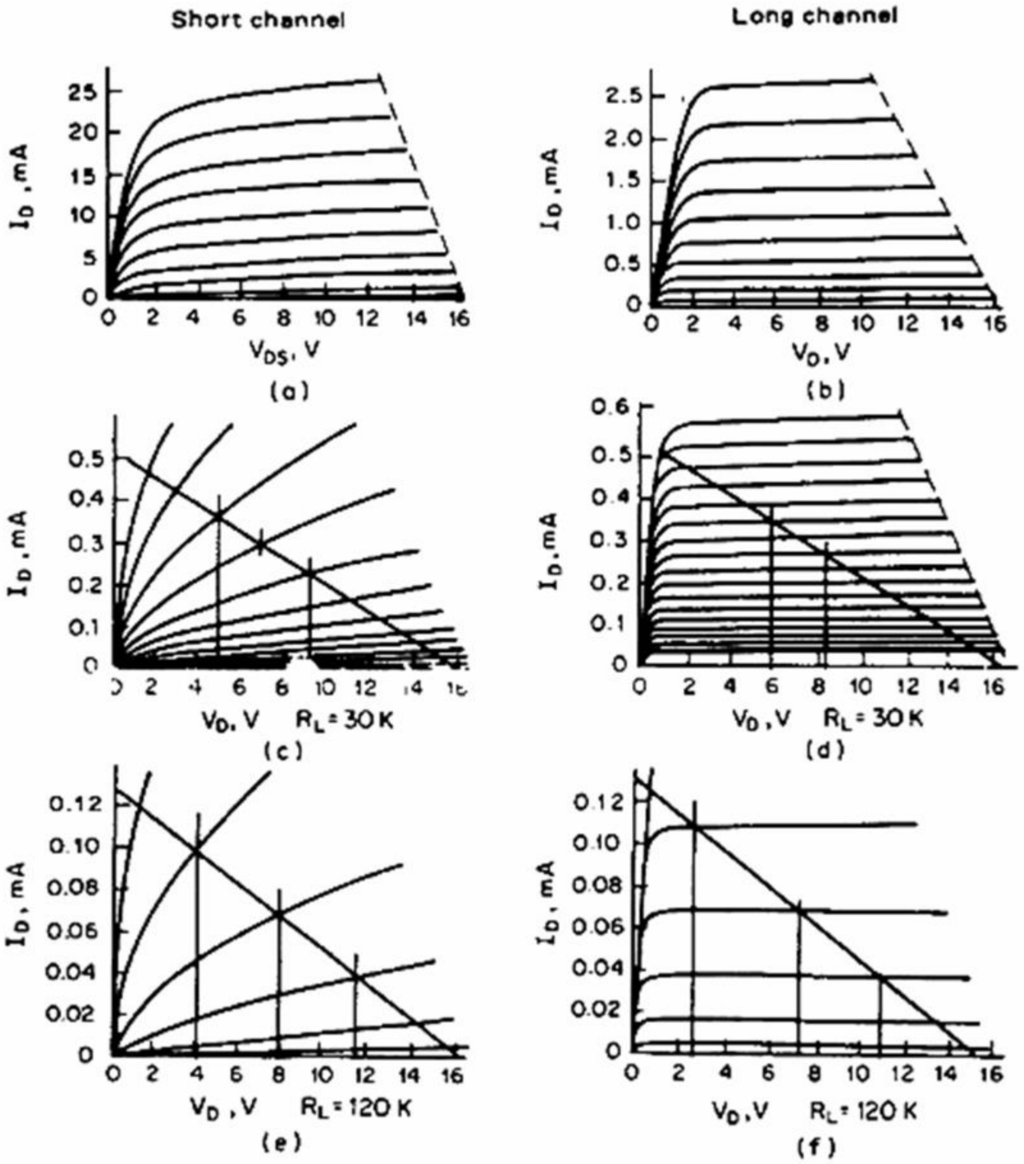

FIGURE 3-5 Comparison of short-channel JFET with long-channel JFET.

Designing with Field-Effect Transistors, Siliconix

Fig.1 From: *Designing with Field-Effect Transistors, 2nd edition, Siliconix Inc, Ed Oxner*

Design of an Armstrong-Type Regenerative Receiver Tunable from 3MHz to 12MHz

This section will report the design, construction and testing of a receiver for the SW bands with the following targets:

1. Must use the Armstrong topology.
2. Must be tunable from 3MHz up to 12MHz.
3. May use salvaged not-easy-to-find parts.
4. Must be of simple conception, sensitive and selective.
5. Must have smooth regeneration control.
6. Designed for headphone/earphone listening.
7. Will operate from a 3-volt battery supply.
8. Must have low parts count and may use only BJTs as active devices.
9. Must be fully operational when built on a solderless breadboard.

The reason for using the Armstrong topology is that the receiver can be designed with little effort and high chances of successful operation. The basic Armstrong receiver consists of an external antenna loosely coupled to a tuned detector-amplifier stage with controlled regenerative gain, followed by an AF amplification chain with an output listening device, an earphone, headphones or loudspeaker. Energy from passing radio waves induce corresponding voltages on the antenna-ground system which are in turn coupled to the input-stage tuning tank by an antenna transformer.

Our regenerative receiver comprises the tuned detector-amplifier stage coupled to an external antenna for input radio signals, followed by two stages of AF amplification with enough output power for driving low-impedance 8-ohm earphones or 32-ohm stereo headphones. Means are provided for control of the regeneration level.

The schematic diagram of the receiver can be seen in Fig.2. First, a description of parts and components will be given, along with some explanation of their importance in the circuit, and following, we will elaborate on the design procedure. The section will end with a study of the RF signal magnification taking place in the detector-amplifier stage as a function of regeneration setting.

The most critical parts to construct or purchase are those of the tuning tank for the desired band. Alternately, they can be salvaged from old or obsolete equipment. The main tuning capacitor, C_1, was found in a spare-parts box and is an air-dielectric variable type with 17pF of minimum capacity and 436pF of maximum capacity. C_2 is a 7pF-17pF air-dielectric variable capacitor, used as a bandspread unit and salvaged from an old vacuum-tube transmitter. It was formerly a larger capacitor, but three plates were separated from the rotor group for lower net capacitance. C_f is a small 6pF-40pF air-dielectric ceramic-insulated variable capacitor and constitutes the manual feedback control. It's also a salvaged part obtained from an old transmitter. Capacitor C_a in series with the input of the receiver is also a salvaged air-dielectric ceramic-insulated 140-pF variable capacitor and is used as a signal attenuator, mostly for SSB signals. For weak signals and for DSBC AM reception it could be desirable to short out the attenuator with a toggle switch. The author found this feature useful, and later added it to the circuit.

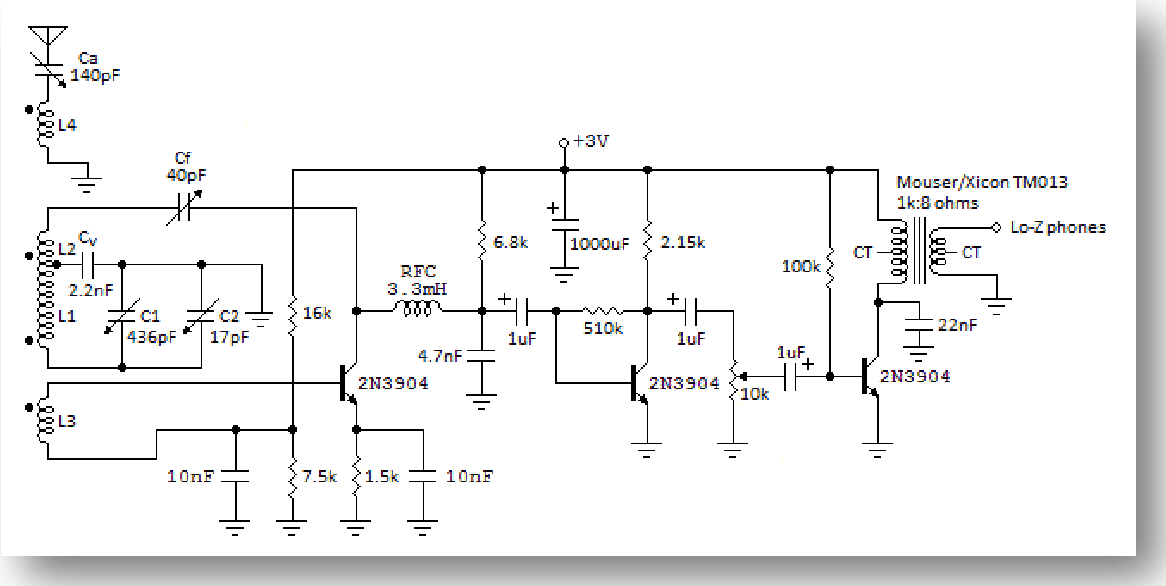

Fig.2 SW Armstrong Regenerative Receiver tunable from 3MHz to 12MHz

C_1 and the paralleled bandspread capacitor C_2 are effectively connected in series with the 2.2-nF Vackar capacitor used for regeneration enhancement throughout the band. They should permit tuning the receiver down to 3MHz when the stray capacitance of the circuit and L_1 are taken into account. For this matter, 7.6uH of inductance will be needed for L_1. However, with C_1 and C_2 at minimum capacitance we will be just below the 25m band and missing a lot of interesting SW listening. For this reason L_1 was left with the number of turns and inductance mentioned in the following paragraph.

The RF transformer consists of windings L_1, L_2, L_3 and L_4, wound on a black-plastic 35-mm film container, as a 3.15-cm diameter coil former. Coils were wound using AWG #25~27 plastic-insulated solid-copper hook-up wire with an equivalent AWG #20 gauge for close-wound solenoids. Design techniques recommend that L_2 should have 1/3 the number of turns of L_1 and L_4 1/6 of that number of turns. L_1+L_2 is a single coil having 17 close-wound turns, with a tap at the 4th turn. L_1 is then 13 turns and L_2 is 4 turns. L_3 is a close-wound 4-turn coil with a 3-mm separation from L_2, measured between wire centers. L_4 is the antenna coil and consists of 2 close-wound turns and has a 6-mm separation from L_1. Using a digital *BK Precision 875A LCR Meter* values for the coils were obtained as L_1 = 6.3uH, L_2 = L_3 = 0.6uH. L_4 was not measured. L_1 was later confirmed to be 6.72uH by a more precise measurement method. Minimum tuned frequency was 3.19MHz. A trade-off had to be made if we wished to tune the 25m band. L_3 permits a better match between the low base-emitter impedance of the 2N3904 transistor from the RF stage and the L_1-C_1 tuning tank, thus preserving selectivity. L_2 is the feedback coil, known also as the tickler, and is the mains for returning a controlled amount of RF energy from the RF stage output back to L_1 in phase with its own magnetic field. The relative phase of the windings is as indicated by the dots in the drawing. All coils are wound in the same direction, L_4 on top, followed by L_1+L_2, then L_3.

As receiving SSB signals is in our plans, we need very stable operation of the detector-amplifier stage in the oscillating mode. A good biasing scheme for this effect is that which uses a two-resistor voltage-divider network for the transistor’s base bias, and an emitter resistor for DC feedback. The 16k-ohm, 7.5k-ohm and 1.5k-ohm resistors fulfill the requirements for the stage bias. The 10-nF capacitors connected in parallel with the 7.5k-ohm and 1.5k-ohm resistors perform RF decoupling.

RF current components at the RF stage output must be kept from flowing into the AF amplifying stages. This is the reason for including the 3.3-mH RF choke (which should be placed at right angles with the input RF transformer in order to avoid unwanted magnetic coupling) and the 4.7nF capacitor in the circuit. They act as a low-pass filter in conjunction with the 6.8k-ohm collector-bias resistor and input resistance of the first AF stage. The AM modulation is recovered across this 6.8k-ohm resistor. The following two AF stages amplify the detected currents to comfortable earphone/headphone listening levels. The 22-nF capacitor connected from the collector of the 2N3904 output transistor to ground is used for tone correction purposes. If the reader is able to purchase from a surplus radio shop a vintage Rochelle-salt piezoelectric high-impedance earphone (these are units having impedances in the 100k-ohm range) or the easier to find low-cost ceramic-piezoelectric 10k-ohm earphone, these may be connected directly across the primary of the output audio transformer. The low-cost ceramic types have been found to fail quite easily if you connect them to a point of the circuit where a DC voltage exists. So, it is much safer to connect them across the primary as said. The vintage Rochelle-salt types don’t seem to present this issue. By the way, the Rochelle types are also known as crystal earphones.

The set of four photos below give a general view of the author’s prototype. In Fig.3.a the brown wires connecting to the receiver at the left are, the antenna lead-in wire, rear most, and the ground lead, foremost. The external ground is a 30cm x 30cm piece of aluminum sheet on the floor next to the desk where the receiver is. This is known as a capacitive ground connection and works very well for SW frequencies.

Design procedure

RF stage

We shall start with the DC design of the detector-amplifier stage. Quiescent-point currents somewhere between 50uA and 200uA have been recommended for a reliable operation of this stage. In early tests, available 2N3904 B331 specimens gave satisfactory results with collector currents in the neighborhood of 200uA, i.e., good sensitivity and smooth regeneration control. This value was adopted in consequence as a design goal. Hereinafter please refer to Fig.4.

Design rules-of-thumb indicate that for the 3-Volt supply, some 10%, or 0.3Volts, should drop across the emitter resistor R_E. Then, R_E = 0.3V/0.2mA = 1.5kohms, assuming as usual, $I_E = I_C$. We are left with 2.7 Volts to distribute among the collector-emitter and collector bias-resistor voltage drops. Accordingly, the voltage drop across R_C is 1.35Volts. Then, R_C = 1.35V/0.2mA = 6.75kohms. We may use the standard value 6.8kohms.

Fig.3.a Front panel with main tuning capacitor, bandspread capacitor and volume control

Fig.3.b Feedback variable capacitor and a 32-ohm pair of headphones in the forefront

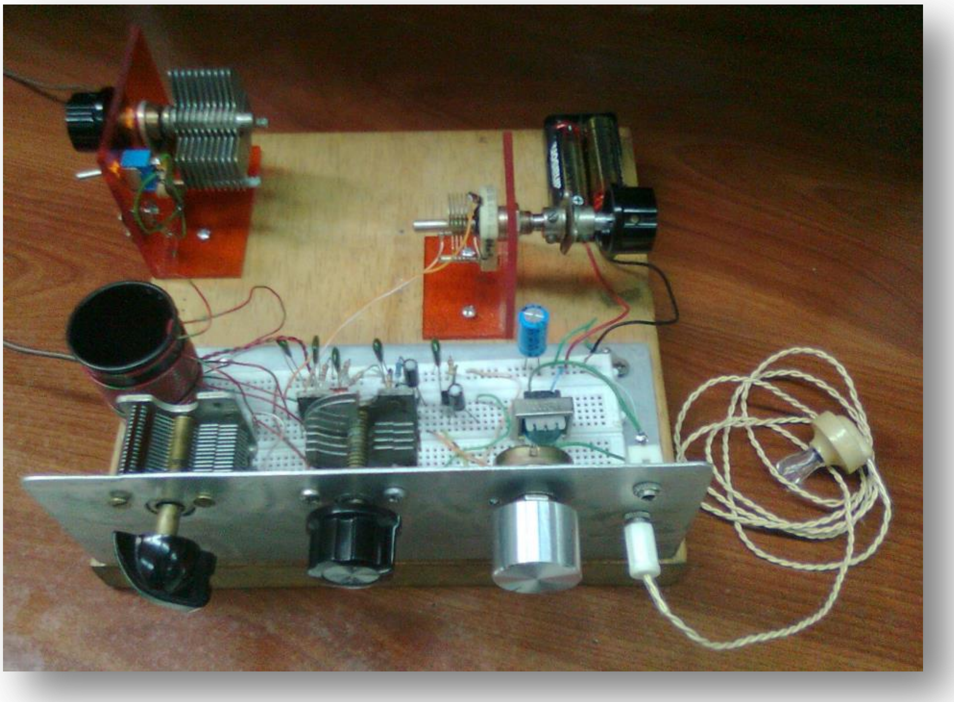

Fig.3.c Capacitive input attenuator at far left and a high-sensitivity ceramic-piezoelectric low-cost 10k-ohm earphone

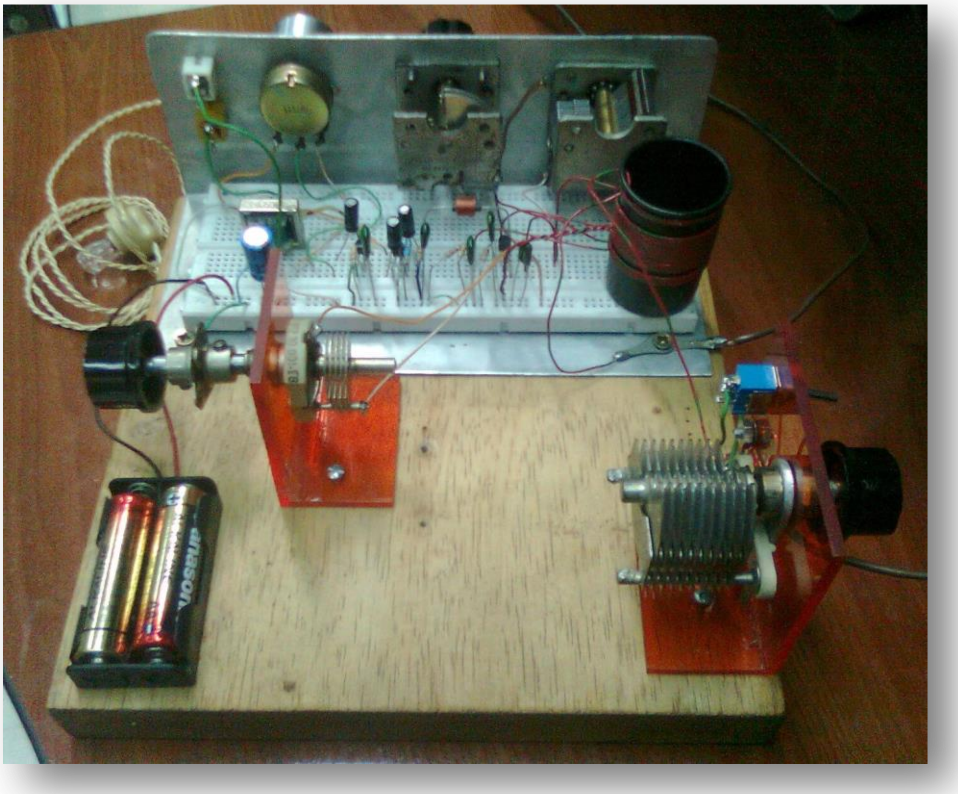

Fig.3.d Rear view of the receiver

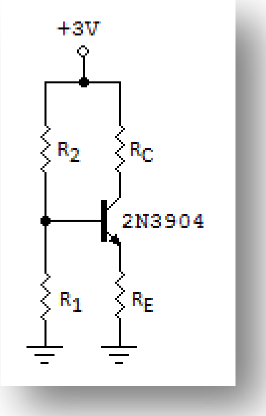

Fig.4 Reduced DC equivalent of detector-amplifier RF stage

Bias equations follow:

$$\frac{V_{CC}R_1}{R_1+R_2}-(R_1//R_2)I_B=V_{BE}+I_ER_E$$

$$V_{CC}\left(\frac{R_1//R_2}{R_2}\right)-\left[\left(\frac{R_1//R_2}{\beta+1}\right)+R_E\right]I_E=V_{BE}$$

If:

$$\frac{R_1//R_2}{\beta+1}\ll R_E \qquad (1)$$

a known condition for emitter-current independence from β, then,

$$\frac{V_{CC}\left(\frac{R_1//R_2}{R_2}\right)-V_{BE}}{R_E}=I_E\approx I_C$$

and

$$V_{CC}\left(\frac{R_1//R_2}{R_2}\right)=I_ER_E+V_{BE} \qquad (2)$$

From Eq.(1), with β = 40 (from 2N3904 datasheet @ 0.1mA) and R_E = 1.5kohms:

$$R_1//R_2\ll(41)(1.5)k\ ohms=61.5kohms$$

$$R_1//R_2\leq 6.15kohms$$

We can adopt $R_1//R_2$ = 5kohms. With V_{CC} = 3Volts and assuming V_{BE} = 0.6Volts, Eq.(2) gives:

$$3\ Volts\cdot\frac{5kohms}{R_2}=0.9\ Volts$$

R_2 is then 16.67kohms. The standard value 16kohms can be used. Substituting in the commonly known expression for $R_1//R_2$, we get 7.14kohms for R_1. A standard value of 7.5kohms will be selected for this resistor.

The values of the decoupling capacitors in parallel with R_1 and R_E are obtained in terms of the maximum permitted reactance as follows:

$$X_{CB} \leq \frac{R_1//R_2}{20} \ @ \ f_{MIN} = 3MHz$$

$$X_{CE} \leq \frac{R_E}{20} \ @ \ f_{MIN} = 3MHz$$

These expressions give the minimum expected values for the capacitors. C_E is selected for maximum audio output with complete absence of unwanted audio-frequency relaxation oscillations, which may appear when the feedback control is advanced. C_B is selected to series resonate with L_3 below f_{MIN}.

Substituting for $R_1//R_2$ we obtain:

$$C_B \geq 212pF$$

This value will null out signals in the neighborhood of 14MHz, as the reader can verify using the known simple resonance formula for L and C in series. A value of 10nF for C_B will produce a series resonance at 2MHz. That would do.

Substituting for R_E in the above respective reactance formula:

$$C_E \geq 707pF$$

Using a 1000-pF capacitor the recovered modulation is weak. Maximum audio output is obtained with 10nF, with no unwanted relaxation oscillations present in the circuit.

Audio stages

The audio-amplifier chain for a regenerative receiver usually consists of two stages capable of giving voltage gains of 1500 or thereabouts, enough for comfortable listening of short-wave stations when using a good external antenna. It is advisable to add a volume control, and in this sense, a potentiometer should be selected having a resistance value of five times the load resistance of the previous stage.

As Fig.2 shows, the needed audio amplification is furnished by a small-signal amplifier followed by a Class-A output stage capable of driving low-impedance audio loads. Both stages are built around general purpose 2N3904's. Parts and components found in a drawer were used for the prototype, bearing in mind the minimum 1500 voltage-gain requirement.

The small-signal audio stage was chosen to use a self-biasing scheme, with collector to base DC feedback and the transistor's collector-emitter voltage drop set at one half of the power supply. A 510k-ohm feedback resistor and a 2.15k-ohm collector resistor yielded the following operating point:

$$V_{CE} = 1.71\,Volts$$

$$I_C = 0.6mA$$

The output stage must deliver enough power for driving 8-ohm earphones or 32-ohm stereo headphones and has to be of simple design. The easiest way to accomplish this is by transformer coupling the low-impedance load to the transistor's collector circuit and biasing the transistor with enough base current for the power required. A 1kohm-to-8ohm XICON 42TM013-RC miniature audio output transformer was used for tests. Fixed base bias was chosen and different resistors in the 100k-ohm range tested for the audio output desired. Trials suggested that DC supply currents as little as 2.5mA could do the work. Accordingly, the base bias resistor was selected to be 100kohms, for the 2N3904 sample employed in the tests. The transformer's primary has a DC resistance of 77 ohms and produces a DC voltage drop of 0.19Volts at a collector current of 2.5mA. So, the operating point of the stage is:

$$V_{CE} = 2.81\,Volts$$

$$I_C = 2.5mA$$

A transistor with different H_{FE} would require a different base resistor value for the same operating point. Other biasing schemes could be tested by the reader for less collector-current dependence on H_{FE} and for thermal stability.

We will now estimate the input resistance of the small-signal audio stage, the voltage gain of this stage, the voltage gain of the output stage and the overall audio voltage gain.

Technical literature has proven elsewhere that our small-signal amplifier has an input impedance in the mid-frequency band, i.e., an input resistance, given by:

$$R_{IN} = \frac{h_{ie}}{1 + h_{fe}\dfrac{R_C//R_L}{R_B}} \tag{3}$$

when $R_B >> R_C//R_L$ and $R_B >> h_{ie}$. More yet, voltage gain is found to be:

$$A_{V_1} = \frac{V_0}{V_B} = -g_m(R_C//R_L)$$

where R_L is the load impedance, in our case, the input resistance of the output stage.

Please refer to Fig.5 for the two expressions above.

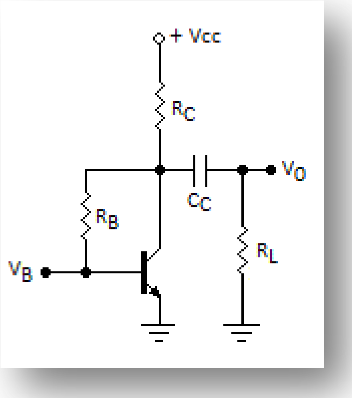

Fig.5 Small-signal amplifier with collector-base DC feedback

The output stage's input resistance is basically the transistor's h_{ie} at the operating point. The input resistance is:

$$R_{IN} = h_{ie} = h_{fe} \cdot \frac{25}{I_{E_Q}}$$

where I_{EQ} is the operating-point collector current in milliamperes. From the 2N3904 datasheet, the h_{fe} is approximately 100 for a 2.5-mA collector current. Then, R_{IN} = 1kohm. The voltage gain is given by:

$$A_{V_2} = -g_m R_L'$$

where $R_L' = N^2 R_L$ is the reflected load impedance as seen by the collector. In our case:

$$N^2 = \frac{1000}{8} = 125$$

from the transformer's datasheet. Then, R_L' = 1kohm for R_L = 8 ohms. The output-stage voltage gain is, numerically,

$$A_{V_2} = -\frac{I_{E_Q}}{25} \cdot R_L' = -100$$

Now A_{V_1} may be expressed by numbers as:

$$A_{V_1} = -\frac{I_{E_Q}}{25}(R_C//R_L) = -\frac{0.6}{25}(2.15//1)(10^3) = 16.38$$

The overall voltage gain of the audio amplifier chain is:

$$A_V = A_{V_1} A_{V_2} = 1638$$

Its input resistance is given by (3):

$$R_{IN} = \frac{h_{fe} \cdot \frac{25}{I_{E_Q}}}{1 + h_{fe} \frac{R_C//R_L}{R_B}} = \frac{\frac{2500}{0.6}}{1 + 100 \cdot \frac{0.6825}{510}} = \frac{4166.67}{1.1338} = 3675\ ohms$$

Study of RF signal magnification in the detector-amplifier stage

With reference to Fig.2, a passing radio wave impresses a voltage signal of the same frequency across the tuning tank formed by L_1, C_1 and Cv in series. Cv is the 2.2-nF Vackar capacitor mentioned elsewhere in this report. Winding L_3 samples this signal and applies it to the base-emitter junction of the transistor for amplification. The amplified version appears between collector and ground and a small amount is fed back to the tuned circuit in phase with the incoming energy. It is our desire to estimate signal-voltage relationships occurring in the stage. Fig.6 shows the tuning tank and feedback network driven by the collector signal V_o. Fig.6.a shows the case when $C_v \rightarrow \infty$, while Fig.6.b is valid for any C_v. The feedback capacitor C_f and the coupling coefficient k_1 between L_1 and L_2 are assumed to have values that permit neglecting the leakage inductance $(1 - k_1^2)L_2$ at the frequencies of interest. Fig.6.c is a good representation for Fig.6.b when the inequality $X_{Cv} << X_{Cf}$ verifies.

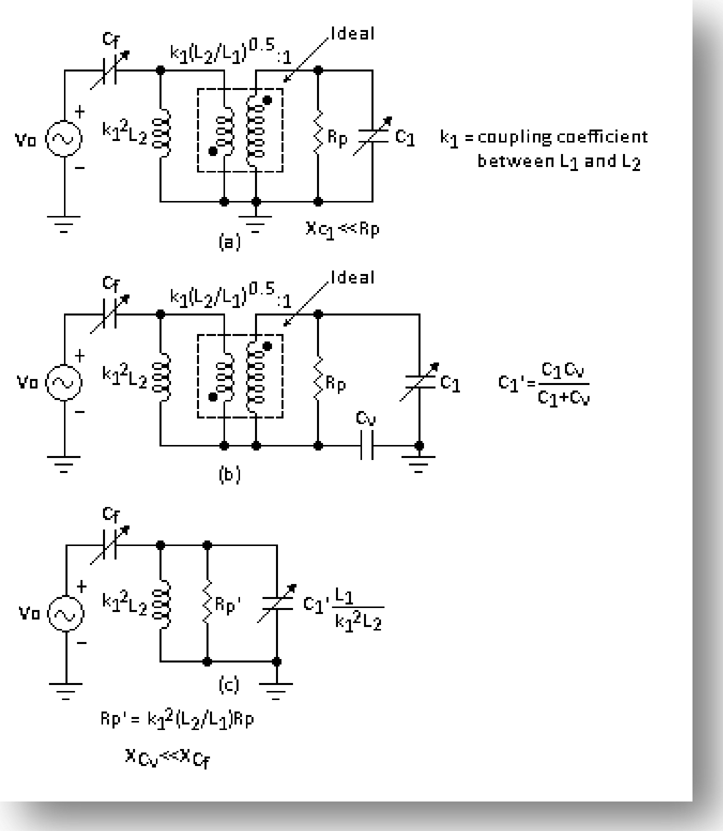

Fig.6 Tuning tank coupled to feedback network and modeled for gain calculations

The input resistance of the detector-amplifier transistor is reflected as a parallel loss across the tuning tank. R_P combines this resistance with the parallel losses of the tank circuit itself.

With reference to Fig.6.b and Fig.6.c, and considering an unmodulated carrier at the resonant frequency ω of L_1 and C_1', the voltage drop across the tickler coil L_2 is:

$$V_P(j\omega) = \frac{j\omega C_f R'_P}{1 + j\omega C_f R'_P} \cdot V_0(j\omega)$$

after noticing that

$$\omega^2 = \frac{1}{L_1 C'_1} = \frac{1}{k_1^2 L_2 \cdot C'_1 \frac{L_1}{k_1^2 L_2}}$$

The voltage across the main tuning coil L_1 is:

$$V_S(j\omega) = \frac{j\omega C_f R'_P}{1 + j\omega C_f R'_P} \cdot \frac{\sqrt{L_1/L_2}}{k_1} \cdot V_0(j\omega)$$

neglecting as a first approximation any voltage contribution due to the interaction of Cv, C_f and the signal source V_o. $V_p(j\omega)$ and $V_s(j\omega)$ are negative at the dotted ends of their respective windings. Then:

$$\begin{aligned} \frac{(-V_S)}{V_0}(j\omega) &= -\frac{j\omega C_f R'_P}{1 + j\omega C_f R'_P} \cdot \frac{\sqrt{L_1/L_2}}{k_1} \\ &= -\frac{j\omega C_f R_P k_1^2 (L_2/L_1)}{1 + j\omega C_f R_P k_1^2 (L_2/L_1)} \cdot \frac{\sqrt{L_1/L_2}}{k_1} \qquad (4) \end{aligned}$$

The base-emitter signal voltage is:

$$V_B(j\omega) = -V_S(j\omega) \cdot k_2 \sqrt{L_3/L_1}$$

where k_2 is the coupling coefficient between L_1 and L_3. The voltage transfer ratio from the tuning tank to the base-emitter circuit is then:

$$\frac{V_B}{(-V_S)}(j\omega) = k_2 \sqrt{L_3/L_1}$$

The base-emitter to collector signal-voltage gain is:

$$\frac{V_0}{V_B}(j\omega) = -g_m \cdot Z_L(j\omega)$$

$$\frac{V_0}{V_B}(j\omega) = -g_m \frac{\frac{1}{j\omega C_f} \cdot R_0}{\frac{1}{j\omega C_f} + R_0} = -g_m \frac{R_0}{1 + j\omega C_f R_0}$$

if $X_{C_f} = 1/\omega C_f \geq 3R'_P = 3R_P k_1^2(L_2/L_1)$. Here, R_o is the transistor's small-signal collector output resistance. The amplifier's loop gain is given by:

$$A\beta = \left(-\frac{V_S}{V_0}(j\omega)\right) \cdot \left(-\frac{V_B}{V_S}(j\omega)\right) \cdot \left(\frac{V_0}{V_B}(j\omega)\right)$$

$$A\beta = \left(-\frac{j\omega C_f R_P k_1^2(L_2/L_1)}{1 + j\omega C_f R_P k_1^2(L_2/L_1)} \cdot \frac{\sqrt{L_1/L_2}}{k_1}\right) \cdot \left(k_2\sqrt{L_3/L_1}\right) \cdot \left(-\frac{g_m R_0}{1 + j\omega C_f R_0}\right)$$

$$= \left(\frac{j\omega C_f R_0}{1 + j\omega C_f R_0}\right) \cdot \left(g_m R_P k_1 k_2 \frac{\sqrt{L_2 L_3}}{L_1}\right) \cdot \left(\frac{1}{1 + j\omega C_f R_P k_1^2(L_2/L_1)}\right)$$

Equating the above expression to *1+j0* will give us the conditions necessary for sinusoidal oscillations to build up. Doing so yields:

$$j\omega C_f R_0 g_m R_P k_1 k_2 \frac{\sqrt{L_2 L_3}}{L_1} = \left(1 + j\omega C_f R_0\right)\left(1 + j\omega C_f R_P k_1^2(L_2/L_1)\right)$$

$$= 1 - \omega^2 C_f^2 R_0 R_P k_1^2(L_2/L_1) + j\omega C_f[R_0 + R_P k_1^2(L_2/L_1)]$$

Then:

$$1 - \omega^2 C_f^2 R_0 R_P k_1^2(L_2/L_1) = 0 \qquad (5)$$

and:

$$g_m R_0 R_P k_1 k_2 \frac{\sqrt{L_2 L_3}}{L_1} = R_0 + R_P k_1^2(L_2/L_1) \qquad (6)$$

Eq.(5) gives the value of the feedback capacitor C_f for which the circuit will be on the threshold of oscillation and maximum amplification of the carrier signal will be obtained. Solving for C_f gives:

$$C_f = \frac{1}{\omega k_1 \sqrt{R_0 R_P (L_2/L_1)}} \qquad (7)$$

where ω is the frequency tuned by the tank circuit. We must be aware that losses in a tuned-network vary with frequency, so *Rp* can't be thought of as a constant.

Eq.(6) is preferably written as:

$$g_m R_P k_1 k_2 \frac{\sqrt{L_2 L_3}}{L_1} = 1 + \frac{R_P}{R_0} k_1^2 (L_2/L_1) \tag{8}$$

and in this form it is known as “the condition for oscillation”.

Now we would like to arrive to an expression for the Vackar-type contribution to *Vs(jω)*. In Fig.6.b we may observe that capacitors C_f and *Cv* are effectively connected in series with the signal source V_o. Following the design constraint given by inequality:

$$X_{C_v} \ll X_{C_f}$$

we can affirm that a voltage:

$$V_v(j\omega) = \frac{C_f/C_v}{1 + j\omega C_f R_P'} \cdot V_0(j\omega)$$

exists across *Cv*. This voltage acts as a signal source for the series circuit formed by L_1 and C_1, as indicated by Fig.7. Polarities are as shown and the reason for not including *Rp* in the equivalent model is that $Rp \gg X_{C_1}$ or equivalently $Rp \gg 1/\omega C_1$ or $Rp \gg \omega L_1$. This is the same as saying that the tuned circuit’s Q is much larger than unity.

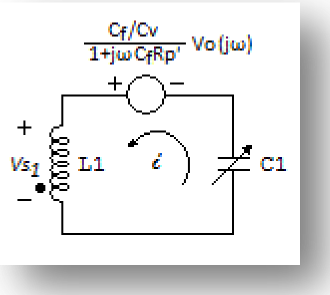

Fig.7 Vackar-type contribution to *Vs(jω)*

In Fig.7:

$$V_{S_1}(j\omega) = \left(\frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}} \right) \cdot \frac{C_f/C_v}{1 + j\omega C_f R_P'} \cdot V_0(j\omega)$$

$$= \left(\frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}} \right) \cdot \frac{C_f/C_v}{1 + j\omega C_f R_P k_1^2 (L_2/L_1)} \cdot V_0(j\omega)$$

Again, ω is the resonant frequency of L_1 and C_1'. For the Vackar-type contribution to appear in the loop-gain expression, Eq.(4) should be replaced by:

$$\frac{(-V_S)}{V_0}(j\omega) = -\frac{j\omega C_f R_P'}{1+j\omega C_f R_P'}\cdot\frac{\sqrt{L_1/L_2}}{k_1} - \left(\frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}}\right)\cdot\frac{C_f/C_v}{1+j\omega C_f R_P'}$$

where:

$$R_P^{'} = R_P k_1^2 (L_2/L_1)$$

Letting:

$$k_0 = \frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}}$$

the loop-gain equation will read as:

$$A\beta = \left(-\frac{j\omega C_f R_P'}{1+j\omega C_f R_P'}\cdot\frac{\sqrt{L_1/L_2}}{k_1} - k_0\frac{C_f/C_v}{1+j\omega C_f R_P'}\right)\cdot\left(k_2\sqrt{L_3/L_1}\right)\cdot\left(-\frac{g_m R_0}{1+j\omega C_f R_0}\right)$$

After equating the expression to $1+j0$ and separating real and imaginary terms we arrive to the following:

$$\omega^2 C_f^2 R_0 R_P k_1^2 (L_2/L_1) = 1 - k_0\frac{C_f}{C_v}k_2\sqrt{L_3/L_1}\,g_m R_0 \qquad (9)$$

and

$$g_m R_P k_1 k_2 \frac{\sqrt{L_2 L_3}}{L_1} = 1 + \frac{R_P}{R_0}k_1^2 (L_2/L_1) \qquad (10)$$

where ω is the tuned frequency. Solving Eq.(9) for C_f will give the value of the feedback capacitor for the threshold of oscillation and maximum RF signal amplification in the circuit. Eq.(10) is the “condition for oscillation” and should be understood as meaning “for the starting of oscillations, the numerical value of the left-hand member must exceed that of the right-hand member”.

Observing Eq.(9), we notice that:

$$k_0\frac{C_f}{C_v}k_2\sqrt{L_3/L_1}\,g_m R_0 < 1$$

must be satisfied in order for C_f to exist. Being $k_2 \leq 1$, a compromise is necessary for the estimation of R_o. Too large a value for the latter and k_2 will be too small for Eq.(10) to verify. R_o is estimated from data available for the RF transistor. The output admittance h_{oe} for the 2N3904 is 4.5µS at Ic = 0.22mA, V_{CE} = 10Volts and f = 1kHz, really not very useful data at HF frequencies. Notice that $R_o = 1/h_{oe}$ = 222.22kohms at said 1kHz. The manufacturer additionally specifies 1µS ≤ $h_{oe} \leq$ 40 µS at Ic = 1mA, V_{CE} = 10Volts and f = 1kHz.

A quick way of checking if the “condition for oscillation” requirement can be satisfied is to see if:

$$g_m R_P k_1 k_2 \frac{\sqrt{L_2 L_3}}{L_1} > 1 \qquad (11)$$

when k_1 is made unity and k_2 a guessed value in the interval 0.1 to 0.4, due to the weak magnetic coupling existing between L_1 and L_3. In practice, k_1 is much closer to unity than k_2. If expression (11) verifies for known (or estimated) Rp, k_1 and k_2, the designer may substitute for these quantities in Eq.(9) and solve for R_o. Remember that L_1, L_2 and L_3 are already known quantities.

The RF choke in the collector circuit of the detector-amplifier transistor contributes with inevitable stray capacitance. A correction factor $1{+}\propto$ must be then introduced in the base-emitter to collector signal-voltage gain expression, which should read as:

$$\frac{V_0}{V_B}(j\omega) = -g_m \frac{R_0}{1 + j\omega C_f (1{+}\propto) R_0}$$

For this reason, $C_f^2(1{+}\propto)$ must substitute for C_f^2 in Eq.(9), and $1{+}\propto$ for unity in the right-hand member of Eq.(10).

The importance of the Vackar-contribution term can be realized checking how much less than unity the right-hand member of Eq.(9) is. The smaller it is, the greater will be the contribution at the frequency of interest.

Here is some experimental data regarding the Vackar-type contribution term to $Vs(j\omega)$. For f = 3.19MHz, the lowest tuned frequency of the receiver, the value measured for C_f was 14.5pF, while Cv measured 2170pF. $C_1' = C_1Cv/(C_1 + Cv)$ was measured as 369.55pF, taking into account 7pF from the bandspread capacitor and 2.4pF of stray capacitance. C_1 was then 436pF+7pF+2.4pF = 445.4pF.

With these data, ωL_1 = 134.69ohms and $1/\omega C_1$ = 112ohms. The Vackar-type contribution term is, accordingly:

$$\left(\frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}}\right) \cdot \frac{C_f}{C_v} = (5.94) \cdot \left(\frac{14.5}{2170}\right) = 0.04$$

For f = 6MHz, C_1 =110.01pF, C_f = 10pF and Cv = 2170pF. We get ωL_1 = 253.34ohms and $1/\omega C_1$ = 241.11ohms. The contribution term is:

$$\left(\frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}}\right) \cdot \frac{C_f}{C_v} = (20.71) \cdot \left(\frac{10}{2170}\right) = 0.095$$

For f = 12MHz, a frequency at the upper end of the tuned band, C_1 = 26.4pF, C_f = 7.2pF and Cv = 2170pF. Now, ωL_1 = 506.67ohms and $1/\omega C_1$ = 502.38ohms. The contribution term is:

$$\left(\frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}}\right) \cdot \frac{C_f}{C_v} = (118.1) \cdot \left(\frac{7.2}{2170}\right) = 0.392$$

For these contribution terms to be applicable, the inequality $R'_P \ll X_{C_f}$ should hold at the tuned frequency of interest. However, the requirement seems to be not so stringent, as the reader can verify that the modulus of $1 + j\omega C_f R'_P$ varies between 1.054 and 1.005 for $3R'_P \leq 1/\omega C_f \leq 10R'_P$.

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Ramón Vargas Patrón
rvargas@inictel-uni.edu.pe
Lima-Peru, South America
May 28th 2015