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Computing RF-Choke Effects in the Loop Gain of an Armstrong-type SW Regenerative Receiver

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Radio experimenters worldwide have a special predilection for the Armstrong topology when designing Medium-Wave Broadcast Band (MW BCB) or Short-Wave (SW) regenerative receivers. This configuration commonly employs a radio-frequency (RF) choke at the output of the detector-amplifier stage to impede RF currents from entering audio-frequency (AF) stages. Popular choke values are 2.5mH and 3.3mH for medium-wave frequencies, and 2.5mH or 1mH for SW reception. Commercial units are manufactured pile-wound or pi-wound on ceramic or phenolic cores. The pi-wound types exhibit very low distributed capacitance and may consist of a single-winding or two or more series-connected sections, a common practice for shifting unwanted parallel resonances outside the operating range of the receiver.

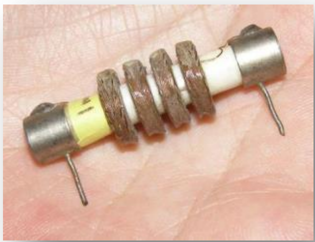


Fig.1 A four-section pi-wound 1-mH RF choke

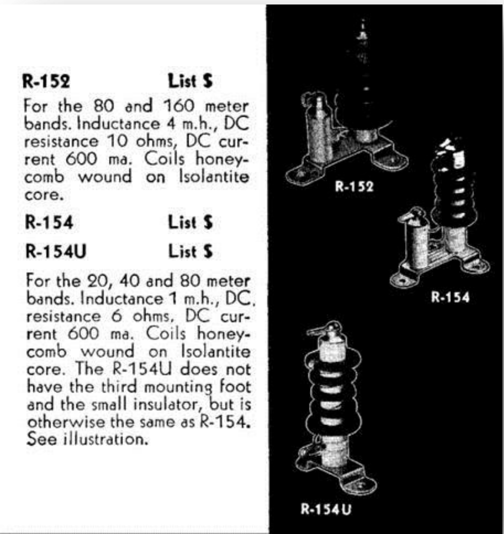


Fig.2 RF chokes for transmitter use

Fig.1 shows a typical pi-wound RF choke and Fig.2 some commercial chokes for RF transmitter applications. An equivalent circuit for a radio-frequency choke can be seen in Fig.3, where R_s represents the DC loss resistance plus the AC loss due to skin and proximity effects. C is the distributed capacitance of the winding and L is the choke's inductance. The model depicts a parallel tuned circuit, which can be of the high-Q or low-Q type, depending on the physical construction of the choke. Usually, low distributed-capacitance values are preferred, which require special winding techniques.

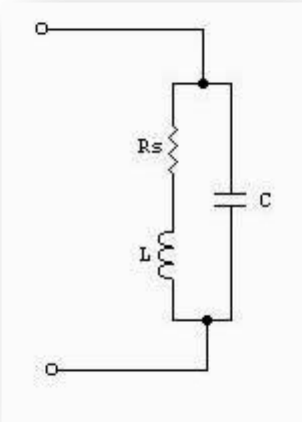


Fig.3 Equivalent circuit for a radio-frequency choke

The Q factor of the coil branch is given by $Q_s = \omega_0 L / R_s$. Losses R_s can be estimated or measured, and so can be the values for Q_s. Let ω_0 be the phase resonant frequency of the network in radians per second. It may be calculated by the expression (Ref.1):

$$\omega_0^2 = \frac{1}{LC}\left(\frac{Q_S^2}{Q_S^2 + 1}\right)$$

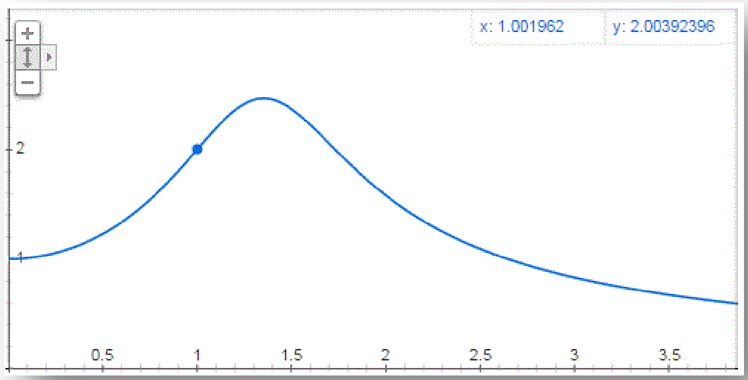
$$= \frac{1}{\left(1 + \frac{1}{Q_S^2}\right) LC}$$

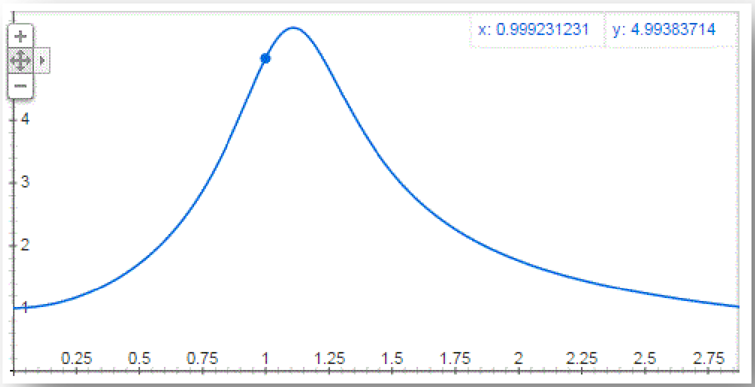
which suggests that an equivalent parallel resonance occurs between capacitor C and an inductor:

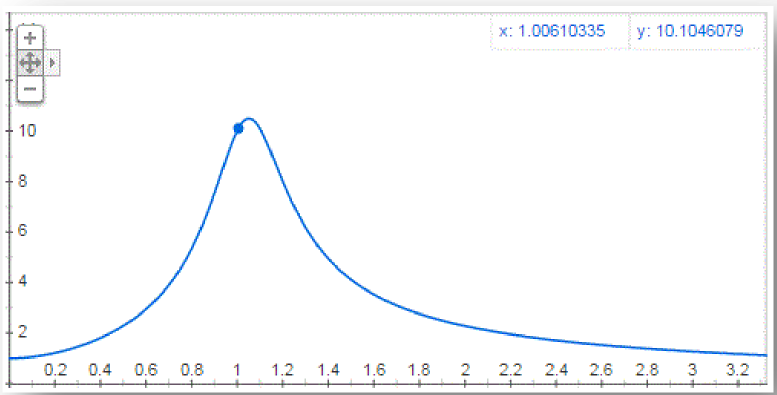
$$L' = L\left(1 + \frac{1}{Q_S^2}\right)$$

Clearly, $\omega_0 < 1/\sqrt{LC}$. Losses have shifted the resonance frequency below the value obtained when ideal reactances are considered in the tuned circuit.

The amplitude response of the network peaks at some frequency *above* ω_0, and the distance in frequency units between the phase resonant frequency ($\omega = \omega_0$) and that corresponding to the peak shortens as Q_s increases. Fig.4 shows normalized impedance amplitude responses for Q_s = 1, 2, 3, 5 and 10.







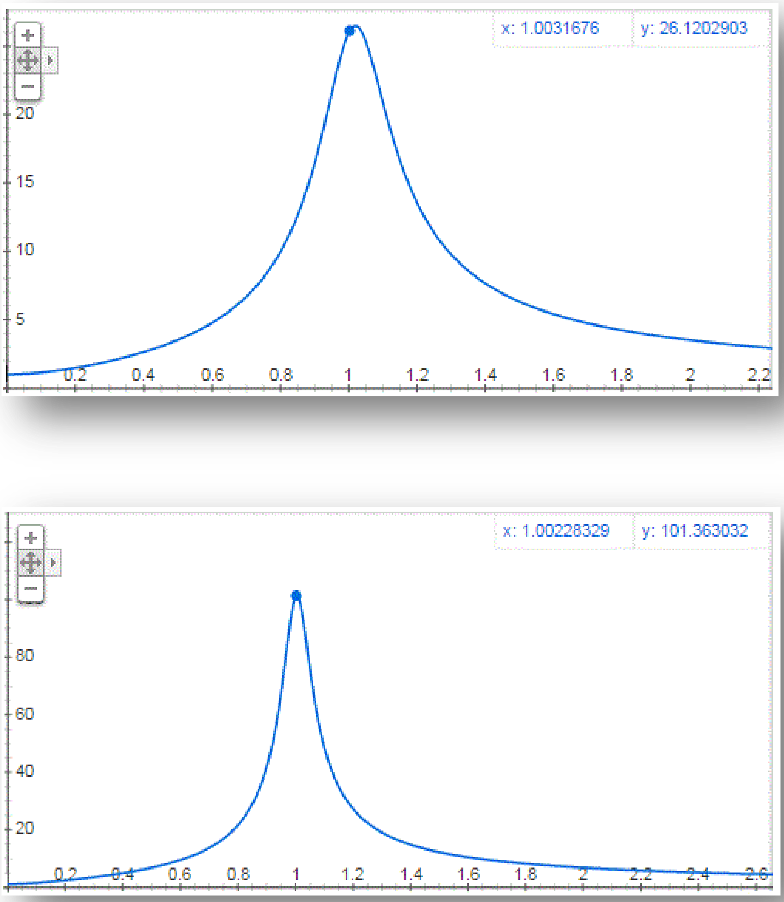


Fig.4 Normalized amplitude response $\dfrac{Z(\omega/\omega_0)}{R_s}$ for Q_s = 1, 2, 3, 5 and 10

The impedance at phase resonance is given by:

$$Z(j\omega_0) = R_P = (Q_s^2 + 1)R_s$$

Below resonance the RF choke presents an inductive behavior. Above resonance, it's predominantly capacitive. The apparent distributed capacitance observed at the network terminals may be computed using the following procedure if the Q factor of the choke is greater than 5. First, we start drawing an RLC parallel tuned circuit equivalent to that shown in Fig.3, and valid for frequencies in the neighborhood of ω_0. Fig.5 shows such a network, which has a resonance peak at ω_0 and a resonant resistance R_P.

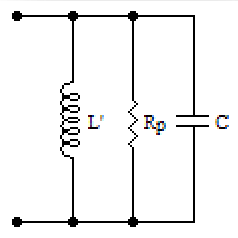


Fig.5 Equivalent circuit for an RF choke when $Q > 5$

The network parameters are:

$$R_P = (Q_s^2 + 1)R_s$$

$$L' = L\left(1 + \frac{1}{Q_s^2}\right)$$

$$\omega_0^2 = \frac{1}{\left(1 + \frac{1}{Q_s^2}\right)LC}$$

Next, we calculate the network admittance as:

$$Y = \frac{1}{R_P} + j\left(\omega C - \frac{1}{\omega L'}\right)$$

$$= \frac{1}{R_P} + j\left(\frac{\omega^2 L' C - 1}{\omega L'}\right)$$

$$= \frac{1}{R_P} + j\left(\frac{\frac{\omega^2}{\omega_0^2} - 1}{\frac{\omega}{\omega_0^2 C}}\right)$$

$$= \frac{1}{R_P} + j\left(\frac{\omega^2 - \omega_0^2}{\omega^2}\right)\omega C$$

$$= \frac{1}{R_P} + j\left(1 - \frac{\omega_0^2}{\omega^2}\right)\omega C$$

This result shows that above resonance the choke behaves as a loss resistor R_P paralleled by an apparent capacitor:

$$C_{app} = \left(1 - \frac{\omega_0^2}{\omega^2}\right) C$$

C_{app} very closely approximates C at frequencies above $4\omega_0$.

In a similar fashion, we can determine the choke's behavior at frequencies below resonance. The network admittance is now written as:

$$Y = \frac{1}{R_P} + j\left(\frac{\omega^2 L' C - 1}{\omega L'}\right)$$

$$= \frac{1}{R_P} + j\left(\frac{\frac{\omega^2}{\omega_0^2} - 1}{\omega L'}\right)$$

$$= \frac{1}{R_P} + \frac{1}{j\omega L'}\left(1 - \frac{\omega^2}{\omega_0^2}\right)$$

$$= \frac{1}{R_P} + \frac{1}{j\omega L'\left(\frac{1}{1 - \omega^2/\omega_0^2}\right)}$$

Below resonance, the choke behaves as a loss resistor R_P paralleled by an apparent inductor:

$$L_{app} = L'\left(\frac{1}{1 - \omega^2/\omega_0^2}\right)$$

Computing RF-choke effects in a practical case

In our article addressed by Ref.2 we discussed the design and construction of a simple solid-state Armstrong regenerative receiver covering the SW band from 3MHz to 12MHz. The little receiver, shown in Fig.6, made use of a 3.3-mH RF choke at the detector-amplifier output as the means for blocking the passage of RF currents into the AF stages, while permitting detected modulation currents to pass through for amplification. The said choke is a commercial unit made by the Bourns/J.W.Miller company, with a Mouser stock number 542-70F33RC. This part features a self-resonant frequency Fo_{min} of 0.8MHz, a quality factor Q_{min} of 52, a DC resistance R_{dc} *max* of 51.6 ohms, and is rated for a DC current I_{dc} *max* of 53mA. It consists of a one-section pi-wound iron-cored coil.

We would like to make an estimation of the distributed capacitance of the choke. From the formula expressing ω_0 in terms of network parameters:

$$\omega_0^2 = \frac{1}{\left(1+\frac{1}{Q_S^2}\right)LC}$$

we obtain, after solving for C:

$$C = \frac{1}{\left(1+\frac{1}{Q_S^2}\right)\omega_0^2 L}$$

Substituting available manufacturer data for the 3.3-mH RF choke, the equation yields $C = 12pF$.

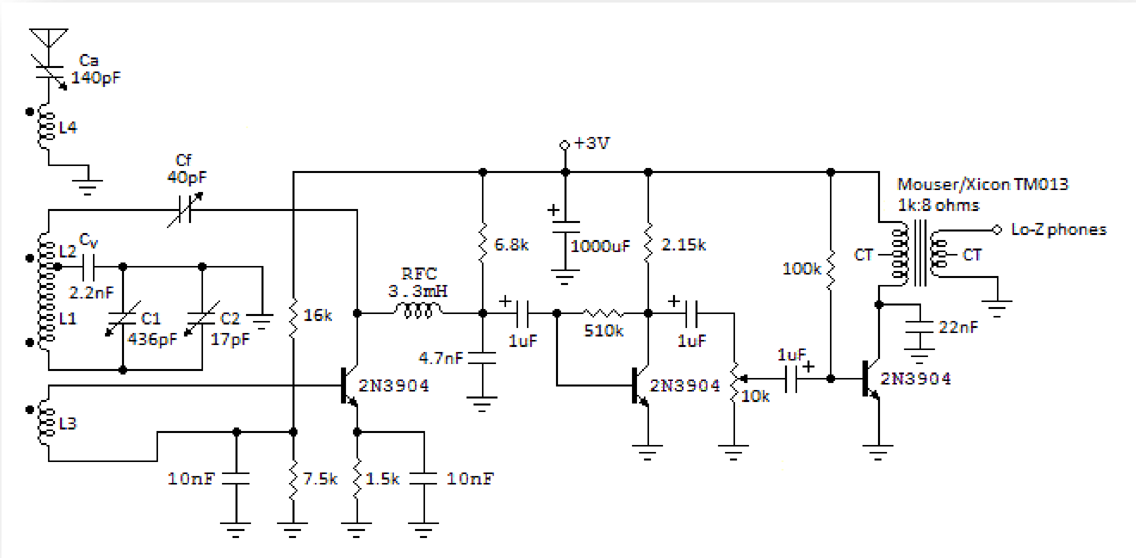


Fig.6 SW Armstrong Regenerative Receiver tunable from 3MHz to 12MHz

In the aforementioned article we arrived to the following expression for the loop gain of the receiver:

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

where:

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

Here, C_1 includes the value of the main tuning capacitor plus that of the bandspread capacitor C_2 for the particular tuned frequency ω. If we recall that $\omega^2 L_1 C_1 = C_1/C_1'$, where $C_1' = C_1 C_v/(C_1 + C_v)$, we can readily write:

$$k_0 = \frac{\omega L_1}{\omega L_1 - \frac{1}{\omega C_1}} = \frac{\omega^2 L_1 C_1}{\omega^2 L_1 C_1 - 1} = \frac{1 + \frac{C_1}{C_v}}{\left(\frac{C_1}{C_v}\right)} = 1 + \frac{C_v}{C_1}$$

This is a preferred expression for k_0, as it will yield smaller computational errors.

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

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

Accordingly, the loop-gain expression equated to $1 + j0$ yields:

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

and

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

Eq.(1) may be expanded to:

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

$$= 1 - C_f \left[\frac{k_0}{C_v} \cdot k_2 \sqrt{L_3/L_1}\, g_m R_0 + \omega^2 C_s R_0 R_P k_1^2 (L_2/L_1) \right] \qquad (3)$$

$$= 1 - C_f \left[\frac{1}{C_1'} \cdot k_2 \sqrt{L_3/L_1}\, g_m R_0 + \omega^2 C_s R_0 R_P k_1^2 (L_2/L_1) \right]$$

$$= 1 - C_f \left[\omega^2 L_1 k_2 \sqrt{L_3/L_1}\, g_m R_0 + \omega^2 C_s R_0 R_P k_1^2 (L_2/L_1) \right]$$

Being $k_0/C_v = 1/C_v + 1/C_1$, the first term within the brackets in Eq.(3) will not vanish when $C_v \to \infty$, because in this case $k_0/C_v \to 1/C_1$. A large value for C_v will still produce some Vackar-like reactive compensation. However, a short circuit between the terminals of this capacitor will null the said term (turns C_v non-existent).

The second term within the brackets highlights the contribution of C_s to overall reactive compensation. It effectively adds to the contribution made by capacitor C_v. On the other hand, capacitance C_s introduces some phase shift in the signal path and influences the value of the transconductance g_m required for maximum signal amplification, which occurs at the threshold of oscillation. This is shown by Eq.(2) above.

Generally speaking, we can distinguish for the receiver the following modes of operation:

1. C_v and C_s in circuit.
2. C_v non-existent and C_s in circuit, when no Vackar-like compensation is required.
3. C_v non-existent and $C_s = 0$, the case when a high-value resistor substitutes for the choke and no Vackar-like compensation is required.

The first mode matches our receiver design (Ref.2). For this mode of operation and any selected frequency in the band, Eq.(3) can be solved for C_f, as long as parameters in the equation remain as known quantities. Tank losses increase with frequency and are usually attributed to coil L_1 (Fig.6). This is a common made assumption when both the main-tuning capacitor C_1 and the bandspread capacitor C_2 are especially selected quality types. As a rough approximation then, $L_1's$ quality factor $Q = \omega L_1 / r$ is assumed to be constant along the tuned band, and if $\mathrm{Q} > 5$, the tank's equivalent parallel loss $\mathrm{R_P}$ will be given by:

$$R_P = Q^2 r = Q\omega L_1 \qquad (4)$$

which also increases with frequency. Any other parallel loss coupled to the tuned circuit will reduce the effective Q value.

Obtaining C_f for the second mode requires re-writing Eq.(3) as:

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

and solving for C_f. Likewise, the required equation for the third mode would be:

$$\omega^2 C_f^2 R_0 R_P k_1^2 (L_2/L_1) = 1 \qquad (6)$$

Following, examples of computed feedback-capacitor values for the three modes of operation will be given for the tuned frequency $f = \omega/2\pi = 12MHz$. Given that coil L_1 was constructed using surplus AWG #25~27 plastic-insulated solid copper wire, a rather low-Q was expected. Measured DC losses for the coil were 2 ohms. By the same token, at 3.19MHz the quality factor can't be greater than $Qmax = \omega L_1 / r_{DC} = 67.34$. If we assume AC losses due to dielectric heating, skin and proximity effects to be equivalent to some 2.5 ohms, the effective coil Q reduces to 30.

The following parameter values will be assumed:
$Q = 30$
$g_m = 0.22mA/25mV = 8.8 \times 10^{-3}\ mhos$
$R_0 = 2000\ ohms$
$C_v = 2170pF$
$C_1 + C_2 = 26.4pF$

$C_s = 12pF$
$L_1 = 6.72uH$
$L_2 = L_3 = 0.6uH$
$k_1 = 0.8$
$k_2 = 0.6$

For the first mode, after substituting Eq.(4) and parameter values in Eq.(3), it reads:

$$0.023\omega^3 C_f^2 + [(2.1202 \times 10^{-5})\omega^2 + (2.76 \times 10^{-13})\omega^3]C_f - 1 = 0$$

Solving for C_f with $\omega = 2\pi \times 12 \times 10^6$ we get $C_f = 3.63\text{pF}$.

For the second mode, after substitution of known quantities Eq.(5) reads:

$$0.023\omega^3 C_f^2 + (2.76 \times 10^{-13})\omega^3 C_f - 1 = 0$$

Solving for C_f and the same frequency we obtain $C_f = 5.72\text{pF}$.

For the third mode, after substitutions Eq.(6) reads:

$$0.023\omega^3 C_f^2 - 1 = 0$$

Solving for C_f and the same frequency it yields $C_f = 10.07\text{pF}$.

Clearly, C_v and C_s work in the same direction introducing reactive compensation in the circuit and reducing the value of the needed feedback capacitor C_f. We have tried to assume realistic parameter values for the examples above. After an open-loop circuit simulation using LTspice IV it seemed correct to assume that R_0 is in the order of a few thousands of ohms. Our receiver uses consumer-grade parts and components, and AC losses have only been guessed, so real-world figures for C_f are expected to largely depart from those given in the above examples. In fact, our receiver working in the first mode required in practice 7.2pF for C_f.

References

1. Vargas Patrón, A. Ramón, *"Design of a Regenerative Receiver for the Short-Wave Bands - A Tutorial and Design Guide for Experimental Work – Part I"*, Technical Report, April 2015, http://www.inictel-uni.edu.pe/publicaciones/reportes-tecnicos
2. Vargas Patrón, A. Ramón, *"Design of a Regenerative Receiver for the Short-Wave Bands - A Tutorial and Design Guide for Experimental Work – Part II"*, Technical Report, May 2015, http://www.inictel-uni.edu.pe/publicaciones/reportes-tecnicos

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