Communication Systems based on Software Defined Radio (SDR)

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Modulators and demodulators of AM

GNU Radio
THE FREE & OPEN SOFTWARE RADIO ECOSYSTEM

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In this class, aspects of the GNU Radio implementation of an AM modulator and demodulator will be discussed.

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The main aim of this practice is to analyze the basic blocks of GNU Radio, the signal analyzers and appreciate how the presence of noise deteriorates the received signal.

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To transmit a signal from one point to another, we must use a carrier signal that is generally of high frequency.

Then, it is possible to modify that carrier based with the information to send. From said carrier signal we can modify:

- AMPLITUDE
- FREQUENCY
- PHASE.

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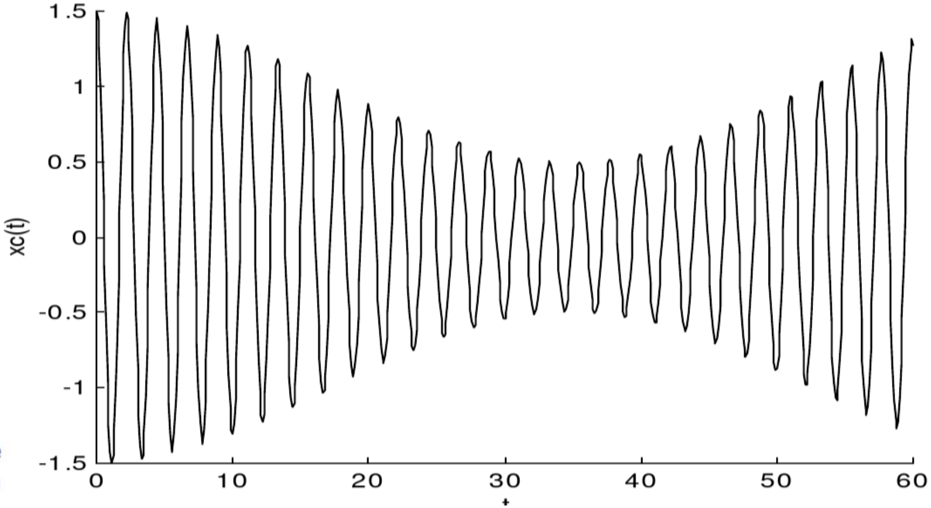
There are three linear modulation methods:

- Amplitude modulation with carrier (AM).
- Double Side Band Modulation (DSB).
- Single Sideband Modulation (SSB or SSB).

In all cases we affect the amplitude of the carrier through the message.



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- Amplitude Modulation with carrier (AM): The envelope of the carrier copies the waveform of the message.

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For mathematical convenience, we normalize our signal so that $|x(t)| \leq 1$. An immediate consequence will be that the average power of the Sx signal will always be less than or equal to one:

$$S_x = \langle x^2(t) \rangle \leq 1 \qquad (5.1)$$

when the signal is deterministic or is a random signal from an ergodic message source. In cases where it is not possible to use the message, we will work with a sinusoidal signal (this case being called as tone modulation) taking as signal:

$$x(t) = A_m \cos \omega_m t \qquad A_m \leq 1 \qquad fm < W \qquad (5.2)$$

Tone modulation makes it possible to simplify power calculations and obtain the output spectrum. For the cases in which it is desired to demonstrate non-linearities, we will work with multi-tone modulation, superimposing several tones, so way to demonstrate potential non-linear effects at different frequencies. In such a case, the sum of the amplitudes of each of the tones must always comply with the normalization condition.

$$x(t) = A_1 \cos \omega_1 t + A_2 \cos \omega_2 t + A_3 \cos \omega_3 t + \ldots \qquad A_1 + A_2 + A_3 + .. \leq 1$$

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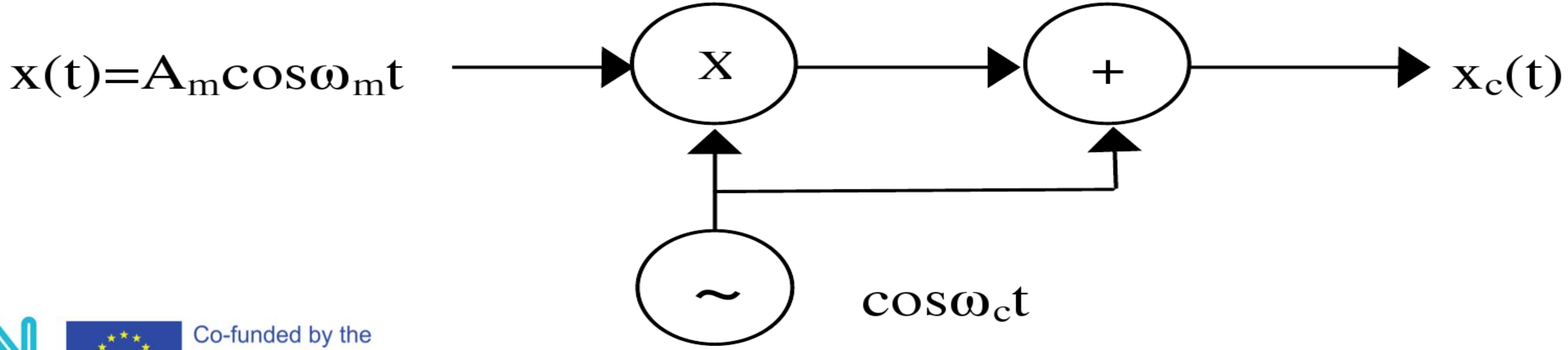
Product modulators

$$x_c(t) = A_c \cos \varpi_c t + \mu x(t) A_c \cos \varpi_c t$$

It is possible to implement the AM equation using the block diagram presented below:



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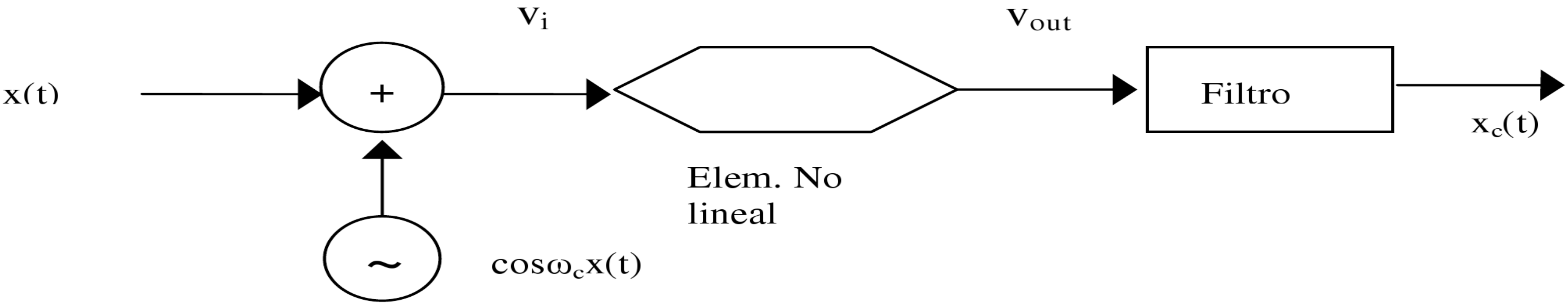
$$x_c(t) = A_c\left[1 + \frac{A_m}{A_c}\cos\varpi_m t\right]\cos\varpi_c t$$

- where:

$$\mu = \frac{A_m}{A_c}$$

Square law modulator



- This type of modulator is based on a non-lineal element.

- Non-lineal elements can be approximated by an second order function:

$$v_{out} = a_1 v_{in} + a_2 {v_{in}}^2$$

- Then, if $V_{IN}(t)$= x(t) + coswct :

$$v_{out}(t) = a_1 x(t) + a_2 x^2(t) + a_2 \cos^2 \varpi_c t + \boxed{a_1\left[1 + \frac{2a_2}{a_1} x(t)\right]\cos \varpi_c t}$$

- The circled term has the form of an AM modulation, with $u=2a_2/a_1$ and $Ac=a_1$. But, the other terms must be removed using a filter.

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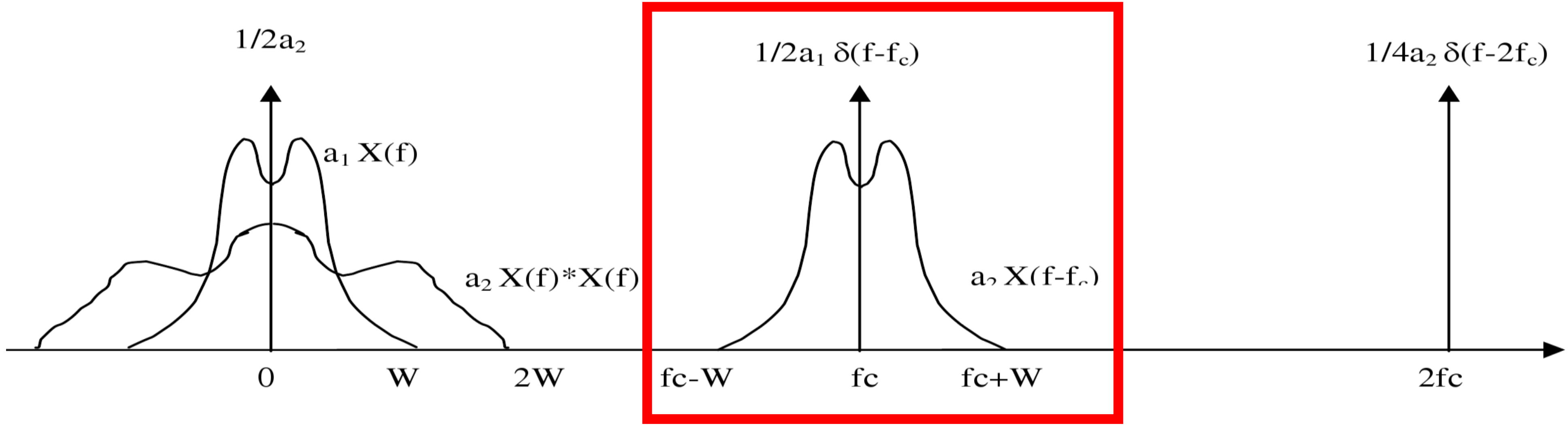
- Grafically:



- The desired AM signal can be obtained using a band pass filter centered in fc, and with bandwith 2W.

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- A propierty of the AM is that the envelope A(t) is similar to the message x(t)

$$A(t) = A_c\left[1 + \mu x(t)\right]$$

- Where μ is a postive constante called modulation index.

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- Then:

$$x_c(t) = A_c\left[1 + \mu x(t)\right]\cos\varpi_c t$$
$$= A_c \cos\varpi_c t + A_c \mu x(t)\cos\varpi_c t$$

And:

$$x_{ci}(t) = A(t) \qquad\qquad x_{cq}(t) = 0$$

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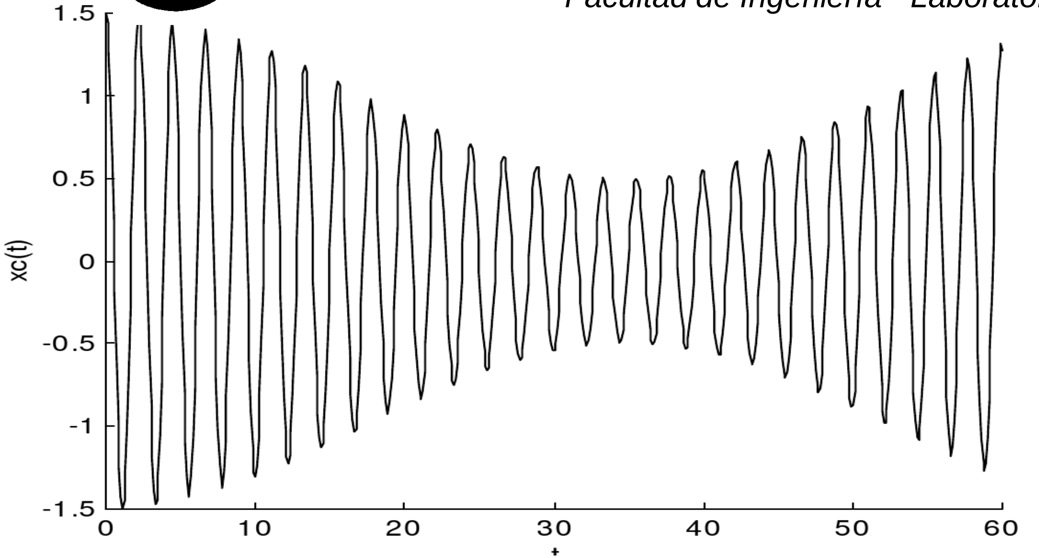

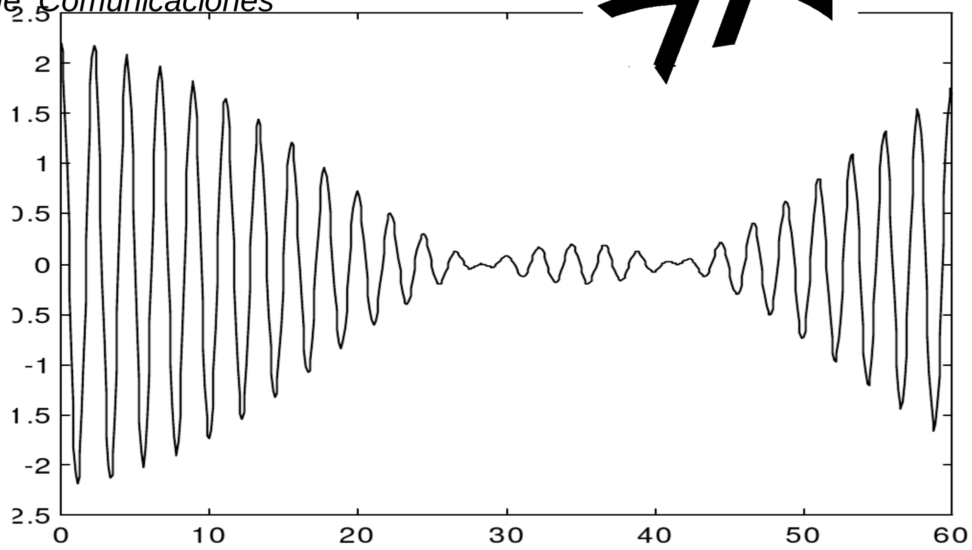

- The envelope shape can be copied if:

$$f_c >> W \qquad \mu \leq 1$$

And, the complete equation of a AM signal is:

$$x_c(t) = A_c\left[1 + \mu \cos \varpi_m t\right]\cos \varpi_c t$$

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- Thus, starting from:

$$\begin{aligned} x_c(t) &= A_c\left[1+\mu x(t)\right]\cos\varpi_c t \\ &= A_c \cos\varpi_c t + A_c \mu x(t)\cos\varpi_c t \end{aligned}$$

- the spectrum of the modulated signal can be obtained:

$$X_c(f) = \frac{1}{2}A_c\delta(f-f_c) + \frac{\mu}{2}A_c X(f-f_c) \qquad f \geq 0$$

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$$X_c(f) = \frac{1}{2} A_c \delta(f - f_c) + \frac{\mu}{2} A_c X(f - f_c) \qquad f \geq 0$$

Graphically:

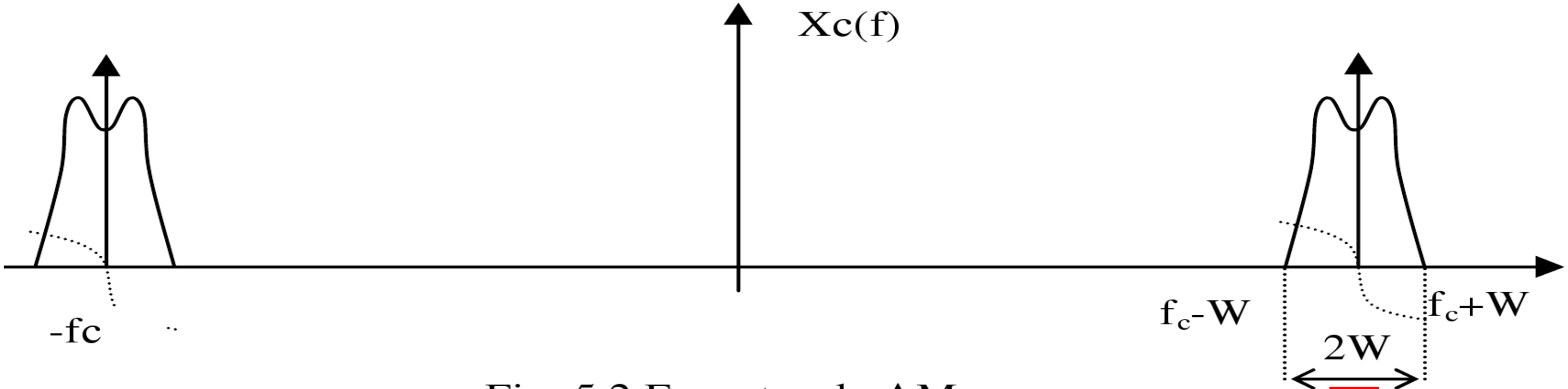

Fig. 5.2 Espectro de AM

$$B_T = 2W$$

- Can be written as:

$$S_T = \langle x_c^2(t) \rangle = \frac{A_c^2}{2}\left(1 + \mu^2 \langle x^2(t) \rangle\right) = \frac{A_c^2}{2}\left(1 + \mu^2 S_x\right)$$

$$S_T = \boxed{P_c} + \boxed{2P_{SB}} = \boxed{\frac{A_c^2}{2}} + \boxed{2\frac{A_c^2}{4}\mu^2 S_x}$$

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"Wasted" power on the carrier can be eliminated by making μ=1 and removing the carrier component that has no modulation. The resulting waveform will be:

$$x_c = A_c x(t) \quad \cos \overline{w}_c t$$

Which is called suppressed-carrier, double-sideband, or more simply double-sideband (DSB). The transform of Equation 5,18 leads to:

$$X(f) \quad = \frac{1}{2} A_c X(f - fc) \qquad\qquad f \geq 0$$

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Where the product of the cosines has been expanded trigonometrically. A similar expansión leads to the following equation for AM:

$$x_{(t)} = \underbrace{\boxed{A_c \cos \overline{w}_c t}}_{\text{Carrier}} + \underbrace{\boxed{\frac{A_c \mu A_m}{2} \cos(\overline{w}_c + \overline{w}_m)} + \boxed{\frac{A_c \mu A_m}{2} \cos(\overline{w}_c - \overline{w}_m)}}_{\text{Side bands}} \quad (5{,}24)$$

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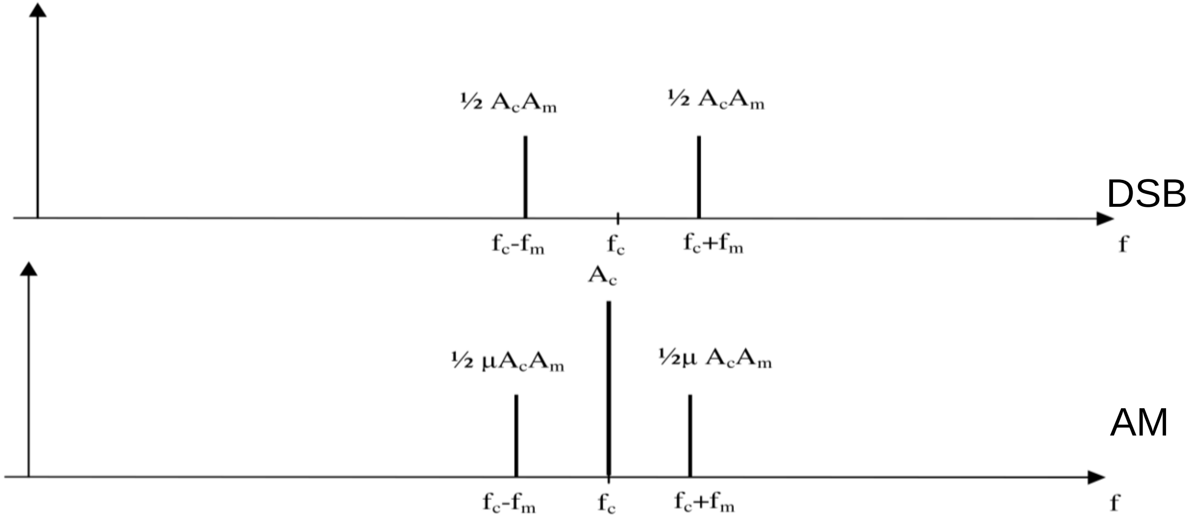
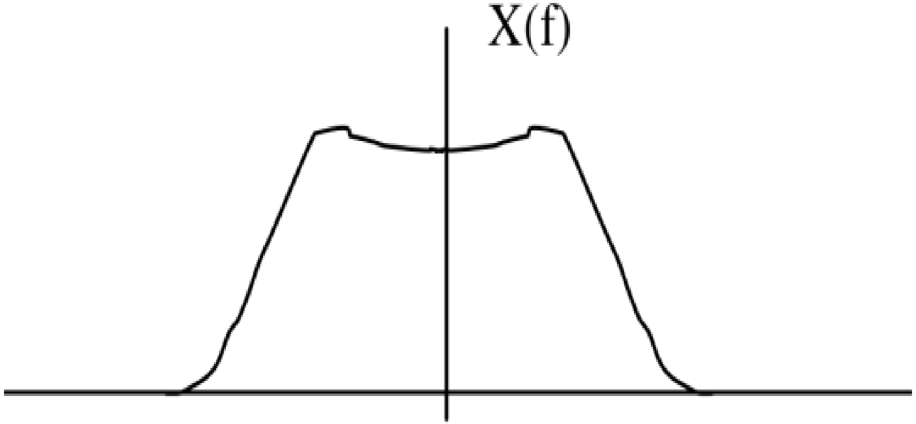
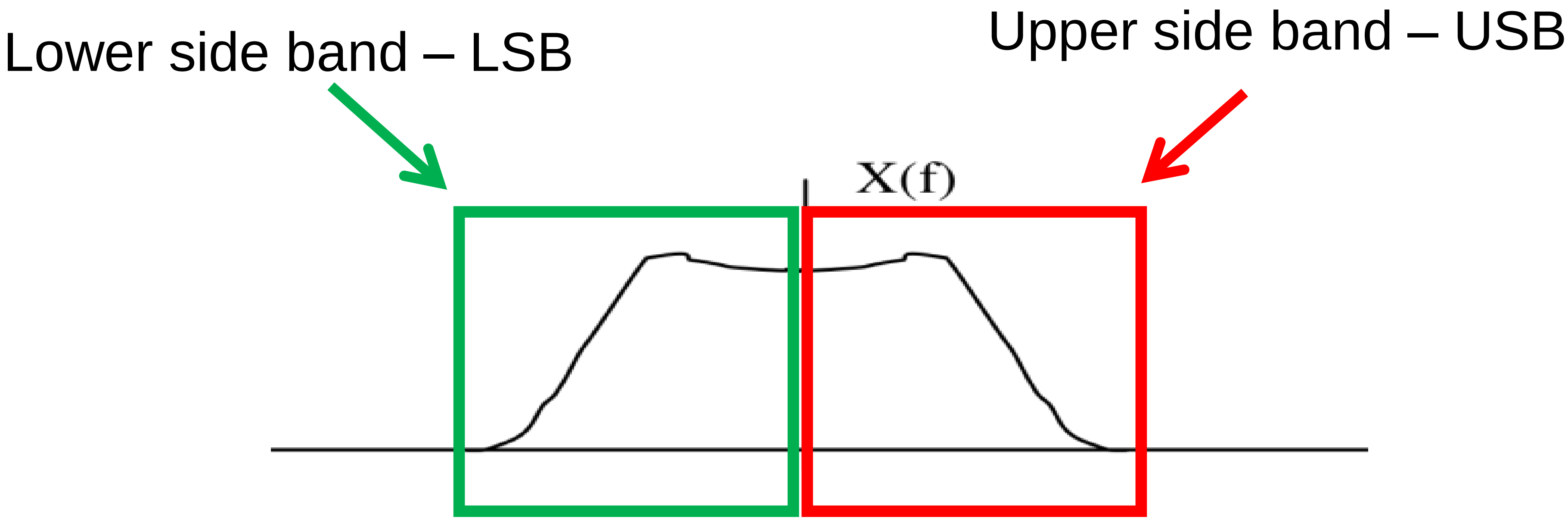
Fig. 5.4-1 shows the spectrum of positive lines obtained with the Eq. 5.23 and 5.24.



- AM wastes both bandwith and transmitted power. Suppressing one of the sidebands produces a method called Single Sideband (SSB). This is based on the fact that each of the sidebands includes all the information of the message.





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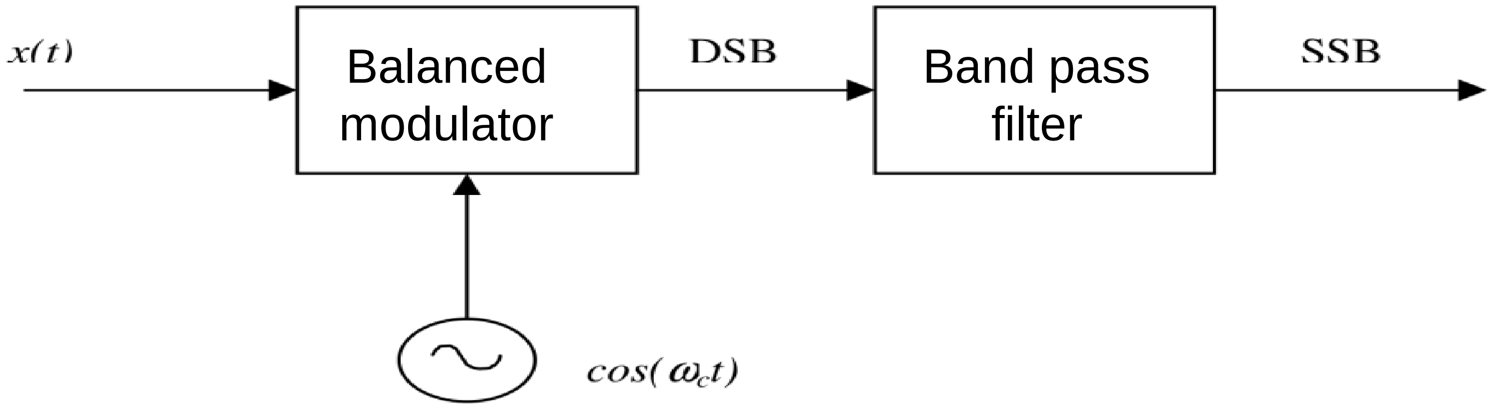
- SSB could be generated starting from a DSB modulator and a filter to remove one of the sidebands.



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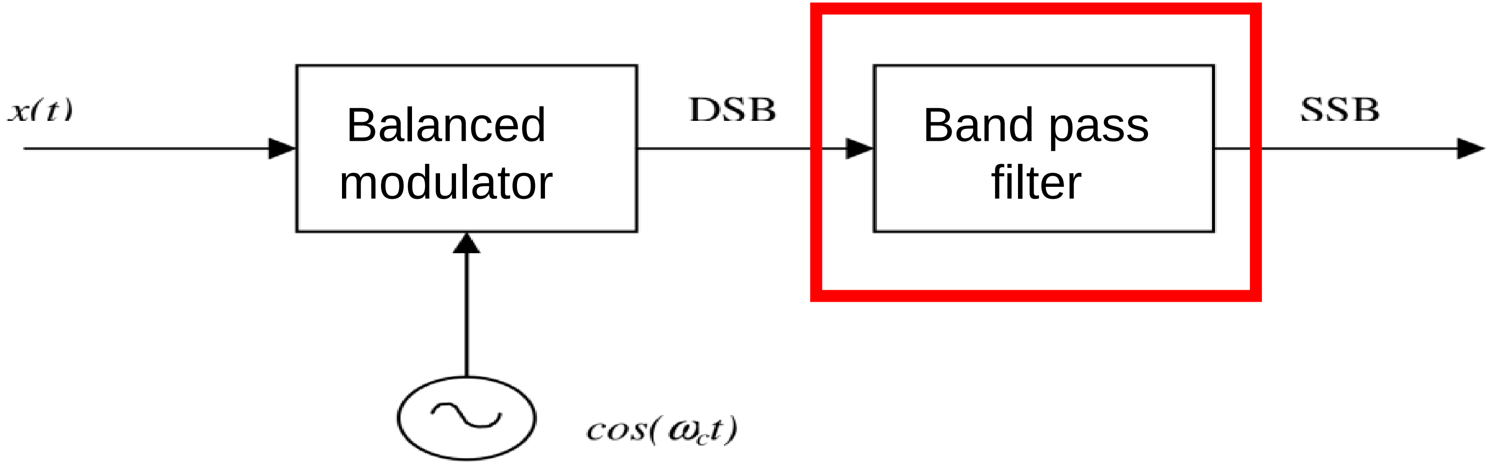
- SSB could be generated starting from a DSB modulator and a filter to remove one of the sidebands.



- The problem with this implementation is that the order requirement of the filter makes it very expensive or even unfeasible.

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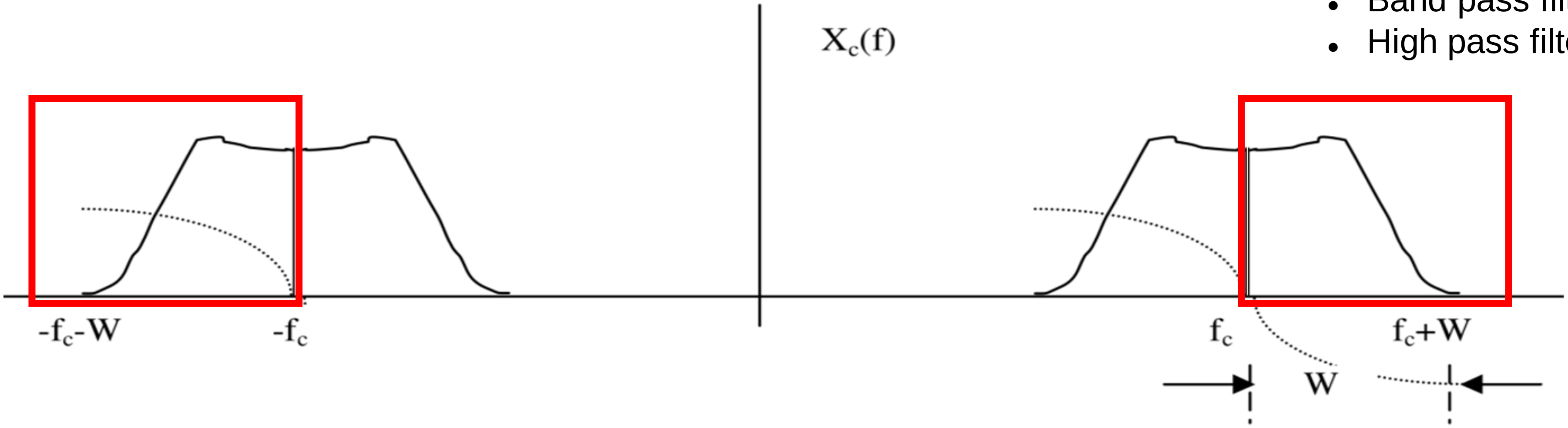
This is because, in order to reject the unwanted band, we must have a filter with a very high slope at a very high working frequency.

- Band pass filter
- High pass filter



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This is because, in order to reject the unwanted band, we must have a filter with a very high slope at a very high working frequency.

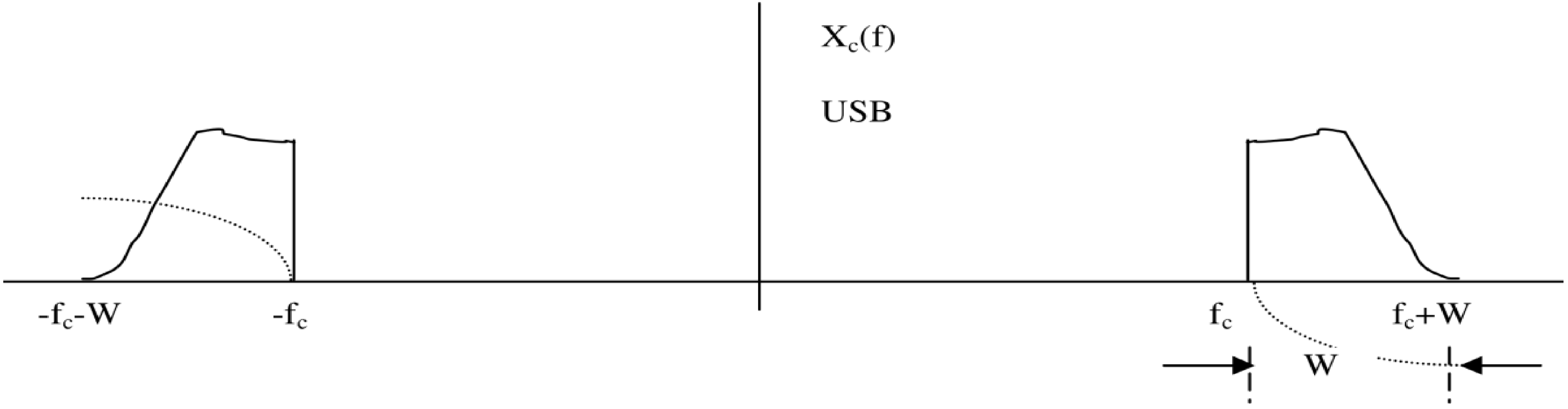

Fig 6.4.3 Spectrum of the upper sideband signal.

- Analogously, for LSB:

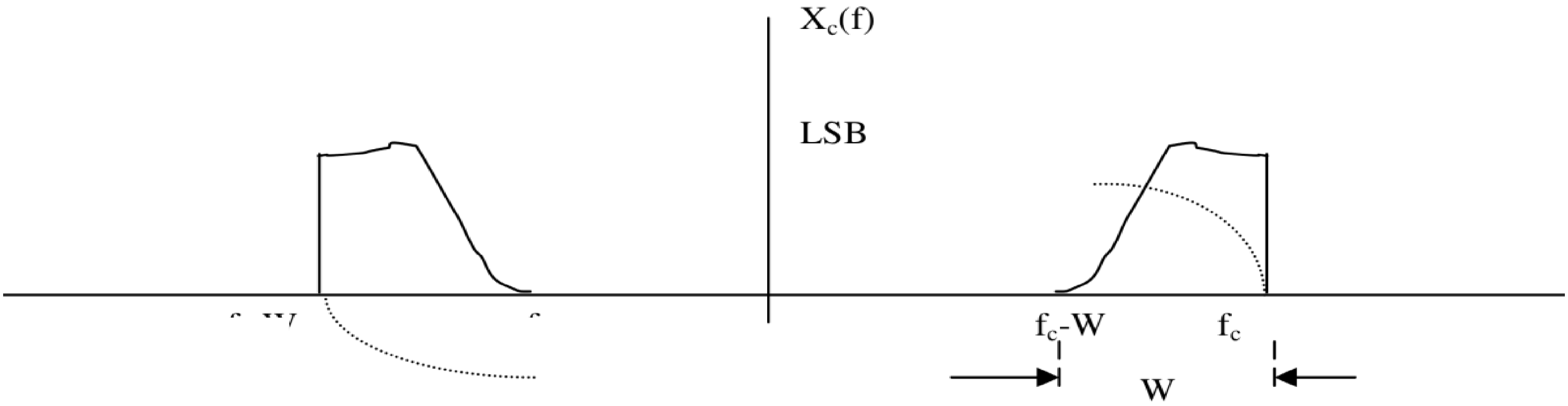

Fig 6.4.3 Spectrum of the upper sideband signal.

- Ideally, depending on the chosen filter, DSB will be transformed into USB or LSB. In any case:

$$B_T = W \qquad\qquad S_T = P_{SB} = \frac{A_c^{\,2}}{4} S_x$$

- In the time domain, it is not so simple to appreciate SSB:

$$x_c(t) = A_c A_m \cos\omega_m t . \cos\omega_c t$$

$$x_c(t) = \frac{A_c A_m}{2}\cos(\omega_c \pm \omega_m)t$$

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$$x_c(t) = \frac{A_c A_m}{2} \cos(\omega_c \pm \omega_m)t$$

- The double sign convention is used, the upper one (+) to indicate USB and the lower one (-) for LSB.

$$x_c(t) = \frac{A_c}{2}[x(t)\cos\omega_c t - \pm\hat{x}(t)sen\omega_c t]$$

- The sign (-) corresponds to USB and the (+) to LSB.

- It can be seen that xc(t) has components in phase and in quadrature:

$$x_{ci}(t) = \frac{1}{2}A_c x(t) \qquad x_{cq}(t) = \frac{1}{2}A_c \hat{x}(t)$$

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$$x_{ci}(t) = \frac{1}{2} A_c x(t) \qquad x_{cq}(t) = \frac{1}{2} A_c \hat{x}(t)$$

- And:

$$A(t) = \frac{1}{2} A_c \sqrt{x^2(t) + \hat{x}^2(t)}$$

- It can be seen that with an envelope detector it is not possible to recover x(t).

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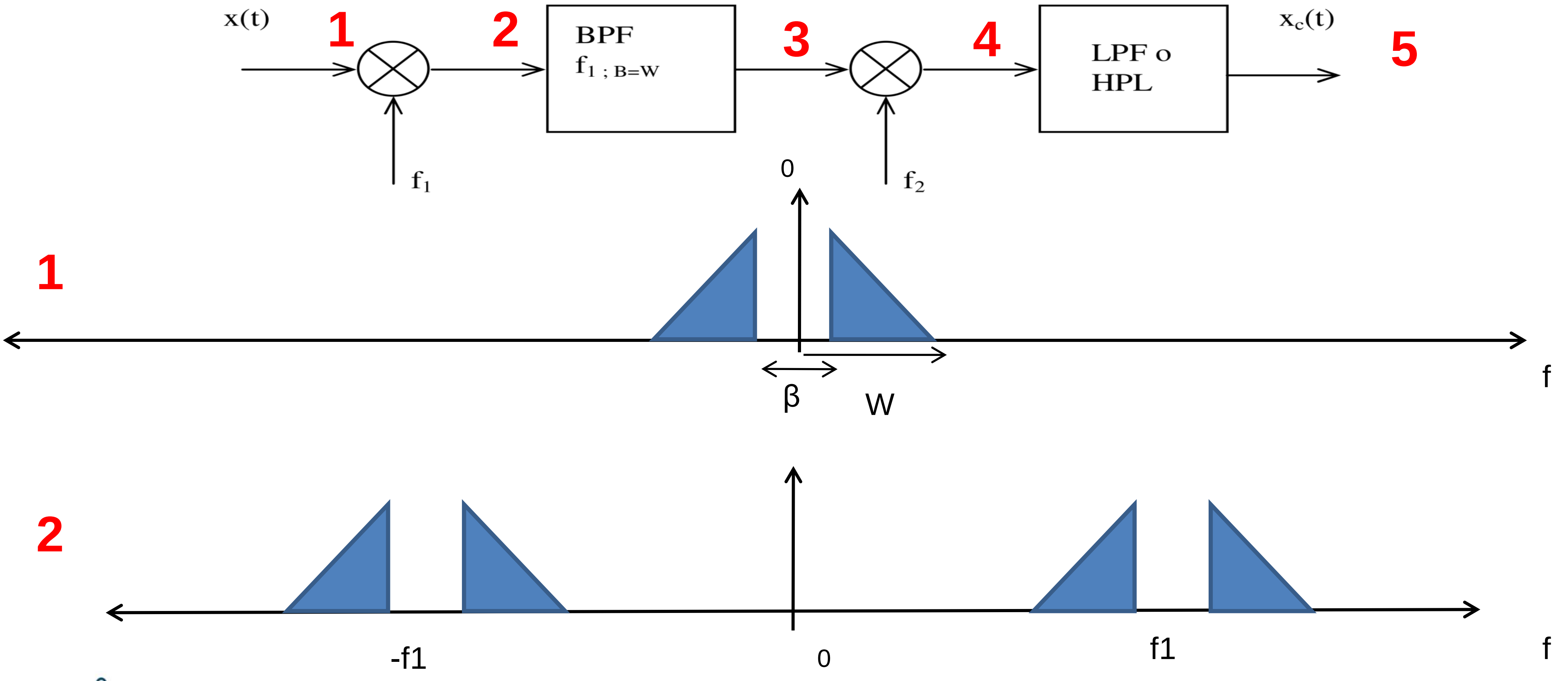
x(t)
1
2
BPF
f1 ; B=W
3
4
LPF o
HPL
xc(t)
5
f1
f2
0
1
β
W
f
2
-f1
0
f1
f

$$x_c(t) = \frac{A_c}{2}x(t)\cos\omega c_c t - \pm\frac{A_c}{2}\hat{x}(t)\cos(\omega c_c t - 90)$$

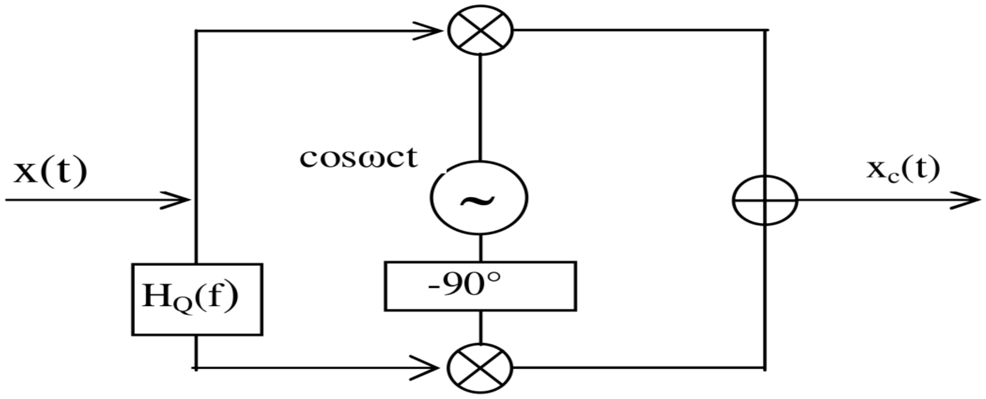

Figure 6.4.11 SSB generation by phase rotation.

This method does not need elaborated filters. The cancellation of one of the sidebands is produced by the phase rotation introduced between the two branches of the circuit. However, the quadrature or rotator filter Hq(f) is a non-realizable network, and can only be roughly synthesized over a narrow frequency range.

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Finally, we are ready to analyze the model presented in Fig. 8.4. The signal xc(t) has modulation linear, and is contaminated with additive white Gaussian noise at the receiver input. The bandpass filtering of predetection produces:

$$v(t) = x_c(t) + n(t) \qquad \overline{x_c^2(t)} = S_R \qquad \overline{n^2(t)} = N_R$$

So the prediction signal to noise ratio is:

$$\left(\frac{S}{N}\right)_R = \frac{S_R}{N_R} = \frac{S_R}{\eta B_T} = \frac{W}{B_T}\gamma$$

Bandpass noise can be expressed by:

$$n(t) = n_i(t)\cos\varpi_c t - n_q(t) sen\varpi_c t$$

Where

$$\overline{n_i^2} = \overline{n_q^2} = \overline{n^2} = N_R = \eta B_T$$

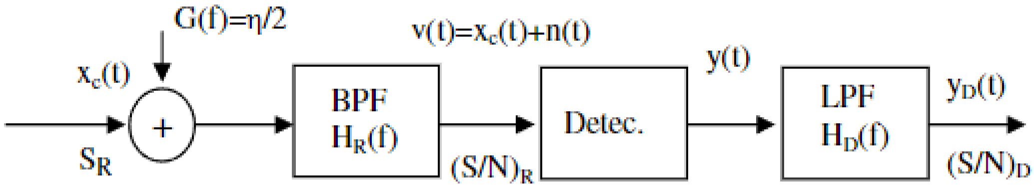

Fig. 8.4 Receiver model for linear modulation with noise.

The question to answer is: given a linear modulation system, what will be the signal to noise ratio in the destination?

Since there are two types of linear detection, synchronous and enveloped, we will look at each of them for the corresponding mode.

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Synchronous detection:

All types of linear modulation can be demodulated by means of a product detector such as that of the Figure 8.5. The input signal is first multiplied by a locally generated sinusoid in the receiver, and then low-pass filtering is performed, where the bandwidth of the filter is equal to that of the message, or slightly elderly. The receiver oscillator, called the local oscillator, is assumed to be exactly synchronized with that of the transmitter, that is, it has the same frequency and the same phase as that of the transmitter. This synchronization gives rise to the name synchronous detector or coherent detector.

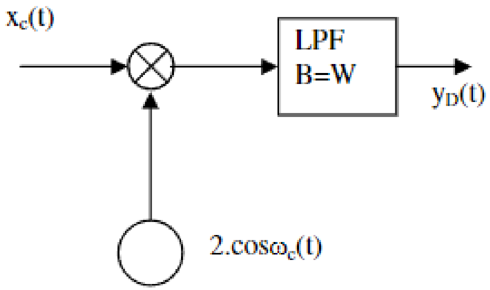

Figure 8.5 Synchronous detector

For the purposes of the analysis we write the input in terms of its components in phase and quadrature, so that can represent all possible modes of linear modulation.

$$x_c(t) = x_{ci}(t).\cos\varpi_c t - x_{cq}(t).sen\varpi_c t \tag{8.28}$$

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For example, depending on the modulation method:

$$\text{AM} \qquad x_{ci}(t) = A_c[1+\mu x(t)] \qquad x_{cq}(t) = 0 \qquad (8.29)$$

$$\text{DSB} \qquad x_{ci}(t) = A_c x(t) \qquad x_{cq}(t) = 0 \qquad (8.30)$$

$$\text{SSB} \qquad x_{ci}(t) = \frac{A_c}{2} x(t) \qquad x_{cq}(t) = \frac{A_c}{2} \hat{x}(t) \qquad (8.31)$$

The input to the filter will be the product of the carrier local oscillator and the input signal:

$$x_c(t).2.\cos\varpi_c t = x_{ci}(t).2.\cos^2\varpi_c t - x_{cq}(t).2.\cos\varpi_c t.sen\varpi_c t \qquad (8.32)$$

Remembering the following trigonometric identities:

$$\cos\alpha.\cos\beta = \frac{1}{2}\cos(\alpha-\beta) + \frac{1}{2}\cos(\alpha+\beta) \qquad sen\alpha.\cos\beta = \frac{1}{2}sen(\alpha-\beta) + \frac{1}{2}sen(\alpha+\beta)$$

Con $\alpha = \beta = \varpi_c t$ and substituting, we get:

$$\boxed{x_c(t).2.\cos\varpi_c t = x_{ci}(t)[1+\cos 2\varpi_c t] - x_{cq}(t)[sen 2\varpi_c t]}$$

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So the detected noise power will be:

$$N_D = \overline{n_i^2(t)} = \eta W$$

The signal to noise ratio can now be obtained:

$$\left(\frac{S}{N}\right)_D = \left(\frac{S}{N}\right)_R = \frac{S_R}{\eta W} = \gamma$$

Which shows that SSB has the same performance as a baseband or DSB system, with respect to noise. To summarize the results found in Eq. 8.40, 8.46 and 8.50, we will state the following properties of linear modulation systems, with synchronous detection of the signal with noise:

1- The message and the noise, if they are additive at the input, are additive at the output of the detector.

2- If the predetection noise spectrum is reasonably flat over the transmission bandwidth, the noise spectrum at the output of the detector is essentially constant over the frequency range of the message.

3- In relation to $(S/N)_D$, SSB does not have a particular advantage over DSB. This is because of the properties of coherence of the two sidebands, which compensates for the noise power reduction of SSB.

4- Leaving aside the "wasted" power in AM, all types of linear modulation have the same performance than baseband, based on average transmitted power and power spectral density of noise.

The preceding statements assume ideal or nearly ideal systems, with fixed average power. If the comparisons are based on the peak envelope power, keeping the same PPE value, SSB will give a $(S/N)_D$ 3dB higher than DSB, and 9dB higher than AM.

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Envelope detection and threshold effect

Since AM is normally demodulated with an envelope detector, we must examine how it behaves the envelope detector when noise is present with the signal, and how it differs from the synchronous detector. To the detector input we have (with μ=1):

$$v(t) = A_c\left[1 + x(t)\right]\cos\varpi_c t + n_i(t)\cos\varpi_c t - n_q(t)sen\varpi_c t \tag{8.53}$$

The construction of the phasor in Fig. 8.7 shows that the phase and the shell of the resultant are:

$$A_v(t) = \sqrt{\{A_c\left[1 + x(t)\right] + n_i(t)\}^2 + \left[n_q(t)\right]^2} \tag{8.54}$$

$$\phi_v(t) = \arctan\frac{n_q(t)}{A_c[1 + x(t) + n_i(t)} \tag{8.55}$$

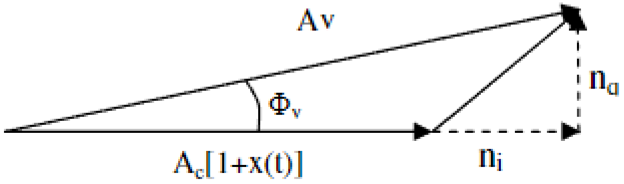

Fig 8.8 Phasor diagram for AM plus noise.

To simplify the expressions of Equations. 8.54 and 8.55, we will make the assumption that the signal is either very large or very small against noise. We will start with the case that the signal is much greater than the noise, that is, that the signal dominates the process detection. In this case, Ac[1+x(t)] is much larger than $n_i(t)$ and $n_q(t)$. In such a case, the envelope can be approximated to:

$$A_v(t) \approx A_c[1 + x(t)] + n_i(t) \qquad (8.56)$$

This shows that an envelope modulation similar to that produced by interference is produced. A ideal envelope detection will reproduce the envelope minus the dc component, so the output of the detector will be:

$$y_D(t) = A_v(t) - \overline{A_v} = A_c x(t) + n_i(t) \qquad (8.57)$$

Which is identical to the output of a synchronous detector. Then, we can use for this case the expression given by the Equation 8.46 for the post-detection S/N. The behavior of the envelope detector is the same as that of the synchronous, so the complexity of synchronization is not justified if we get the same results. But this is only for the case where the signal dominates and not the noise, which translates into the similar requirement:

$$A_c^2 >> \overline{n^2} \qquad \frac{\grave{A}_c^2}{\overline{n^2}} >> 1 \qquad \frac{S_R}{\eta B_T} >> 1 \qquad (8.58)$$

So we can say that envelope detection requires that the predetection signal-to-noise ratio be much greater than one, so that its behavior is equal to the synchronous detector. Without this condition, you cannot express the signal and the noise in an additive way, as stated in Equation. 8.57.

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At the other extreme, when $(S/N)_R << 1$ the situation is totally different. Noise dominates detection in similar to a strong interference, and we can think of it as $x_c(t)$ equals the noise of the previous case. To analyze it, we will express n(t) in the form of envelope and phase:

$$n(t) = A_n(t)\cos[\varpi_c t + \phi_n(t)] \qquad (8.59)$$

The phasor diagram shown in Fig. 8.9 corresponds to this case, since we are taking n(t) as dominant. As done above, the envelope can be approximated as:

$$A_v(t) \approx A_n(t) + A_c[1 + x(t)]\cos\phi_n(t) \qquad (8.60)$$

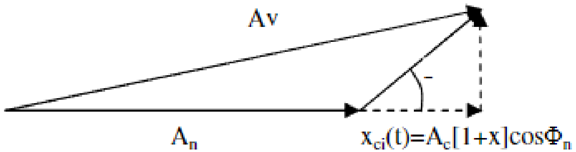

Fig. 8.9 Phasor diagram for AM with noise when $(S/R)_R << 1$

Using this approximation, the detector output will be:

$$y(t) = A_n(t) + A_c x(t)\cos\phi_n(t) - \overline{A_n} \qquad \text{donde} \qquad \overline{A_n} = \sqrt{\frac{\pi N_R}{2}} \qquad (8.61)$$

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The main component of the output is obviously the noise envelope An(t), but more importantly, there is no terms strictly proportional to the message; despite the fact that signal and noise are additive at the input of the detector, the detected message is multiplied by a random function that is $\cos\Phi_n(t)$. Therefore, the message is mutilated, and the information has been lost. Under these circumstances, there is no message detected, and it has no It makes no sense to define a signal-to-noise ratio, since the detected signal output does not exist. The mutilation or loss of the message, from a certain level of signal and noise, is called the threshold effect. The name originates from the fact that from a certain level or value of the (S/N)R ratio, it begins to become mutilation is present and system performance deteriorates rapidly. Actually, it's not a single value, but a range of values, in which the limiting cases do not hold, that $(S/N)_R >> 1$, or that $(S/N)_R << 1$, that is the range in which N_R and S_R are comparable, and mutilation begins. In contrast, with synchronous detection, the The output of the detector is always additive, and even if $(S/N)_R << 1$, the identity of the message is preserved. To set a single value, we will define the inequality Ac >> An. The effects of threshold and mutilation are minimum if Ac >> An for most of the time. More specifically, we define the threshold level, as the value of $(S/N)_R$ for which Ac > An with probability 0.99 The detected noise amplitudes have a Rayleigh distribution as previously indicated, so

$$p(A_c \geq A_n) = 0.99 \qquad p(A_n \geq A_c) = 1 - 0.99 = 0.01 \tag{8.62}$$

$$p(An \geq a) = \int_a^\infty \frac{A_n}{N_R} e^{-\frac{A_n^2}{2N_R}} dA_n = e^{-\frac{a^2}{2N_R}} = 0.01 = e^{-\frac{A_c^2}{2N_R}} \tag{8.63}$$

$$\frac{A_c^2}{4N_R} = \ln 10 \tag{8.64}$$

$$S_R = \frac{1}{2} A_c^2 [1 + u^2 S_x] = A_c^2 \tag{8.65}$$

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We can now define the threshold, in terms of signal power and noise:

$$\left(\frac{S}{N}\right)_{UMB} = 4\ln 10 \approx 10 \qquad (8.66)$$

For AM, the pre-detection signal-to-noise ratio given by Eq. 8.46 with μ=1 and Sx=1, is γ/2, so:

$$\gamma_{UMB} = 8\ln 10 \approx 20 \qquad (8.67)$$

If $(S/N)_R < (S/N)_{THR}$ (or what is the same, $\gamma < \gamma_{THR}$), mutilation will occur and the consequent loss of information. The threshold effect is not a serious limitation in the broadcasting service, since the reception of audio signals imposes a minimum $(S/N)_D$ of 30 dB for reasonable listening. Considering that $(S/N)_D < (S/N)_R$ for In this mode, it is clear that $(S/N)_R$ will be well above the threshold. In other words, additive noise obscures or makes it difficult to perceive the audio signal before the multiplicative noise of the threshold effect mutilates the signal. Therefore, the threshold effect does not constitute a practical limitation in AM modulated audio signals. In others applications, such as digital data transmission, synchronous detection should be used to avoid threshold effect. Finally, let us consider how an envelope detector can act as a synchronous detector, and why. you need a big $(S/N)_R$. In the envelope detector, the diode works like a key that closes very briefly, on the positive peaks of the carrier, so the switching is perfectly synchronized with the carrier. But when noise dominates, switching is primarily controlled by noise spikes, and synchronism is lost. This latter effect never occurs in a synchronous detector, where the carrier locally generated is always much bigger than noise.

- In synchronous detection, although the S/N depends on the noise level, the noise does not garble the message.

- In envelope detection, if $(S/N)_R \gg 1$, the detector performs as well as synchronous. But if $(S/N)_R \ll 1$, the signal is garbled and the message cannot be recovered.

- Although the threshold level establishes that clipping appears if $(S/N)_R < 10dB$. Then, $(S/N)_D < (S/N)_R$ and since with $(S/N)_D < 30dB$ perception is bad, the envelope detector is suitable for broadcasting.

- Therefore, the threshold level for an envelope detector is well below the reasonable listening level.

- **This, added to its low complexity and cost, has contributed to the widespread use of this receiver for broadcasting. In an SDR receiver, the reception will be synchronous, being able to avoid the threshold effect.**

- **In an SDR receiver, the reception will be synchronous, avoiding the threshold effect.**

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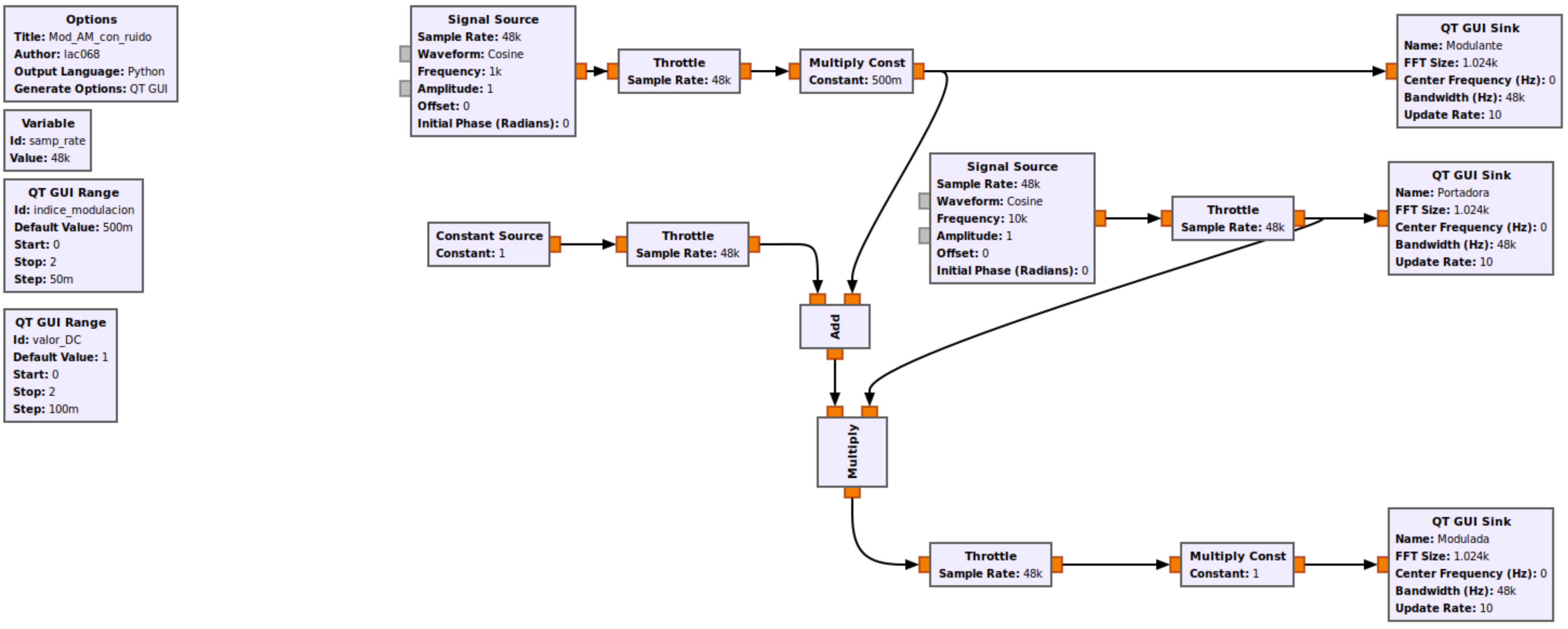
Facultad de Ingeniería - Laboratorio de Comunicaciones



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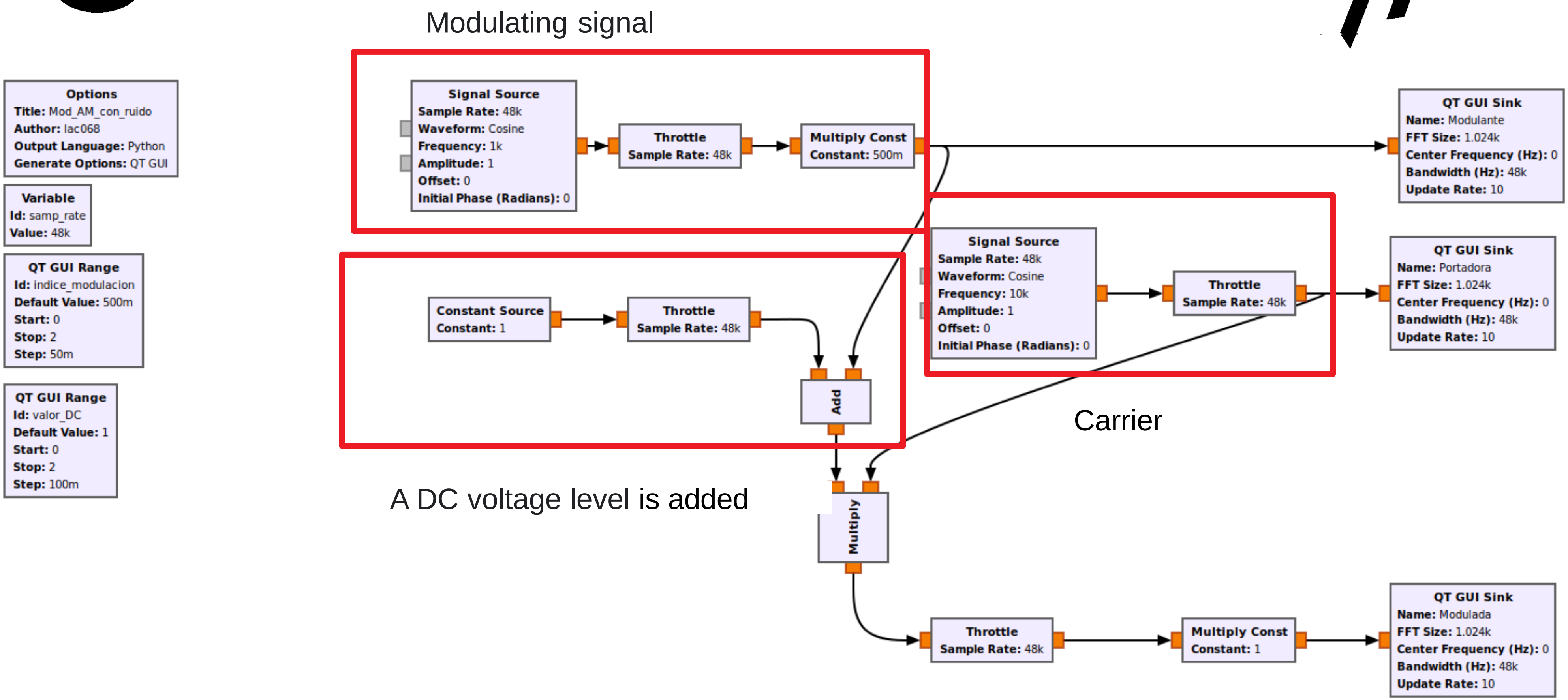

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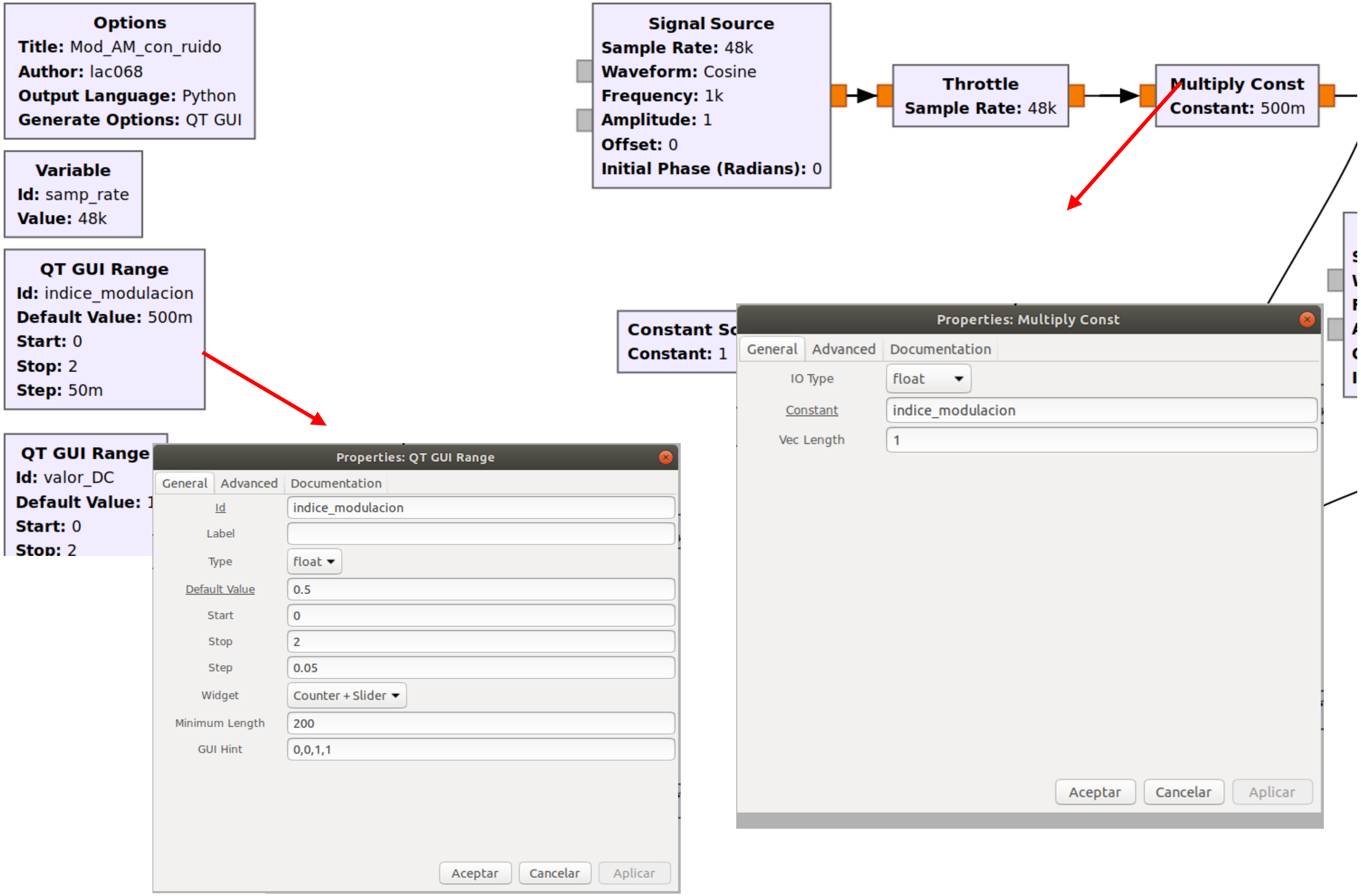
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Title: Mod_AM_con_ruido
Author: lac068
Output Language: Python
Generate Options: QT GUI
Variable
Id: samp_rate
Value: 48k
QT GUI Range
Id: indice_modulacion
Default Value: 500m
Start: 0
Stop: 2
Step: 50m
QT GUI Range
Id: valor_DC
Default Value: 1
Start: 0
Stop: 2
Signal Source
Sample Rate: 48k
Waveform: Cosine
Frequency: 1k
Amplitude: 1
Offset: 0
Initial Phase (Radians): 0
Throttle
Sample Rate: 48k
Multiply Const
Constant: 500m
Constant Sc
Constant: 1
Properties: QT GUI Range
General
Advanced
Documentation
Id
indice_modulacion
Label
Type
float
Default Value
0.5
Start
0
Stop
2
Step
0.05
Widget
Counter + Slider
Minimum Length
200
GUI Hint
0,0,1,1
Aceptar
Cancelar
Aplicar
Properties: Multiply Const
General
Advanced
Documentation
IO Type
float
Constant
indice_modulacion
Vec Length
1
Aceptar
Cancelar
Aplicar

Sliders

Sinusoidal modulating fm=1kHz, m=0,5.

Sinusoidal carrier fc=10kHz

Modulated signal

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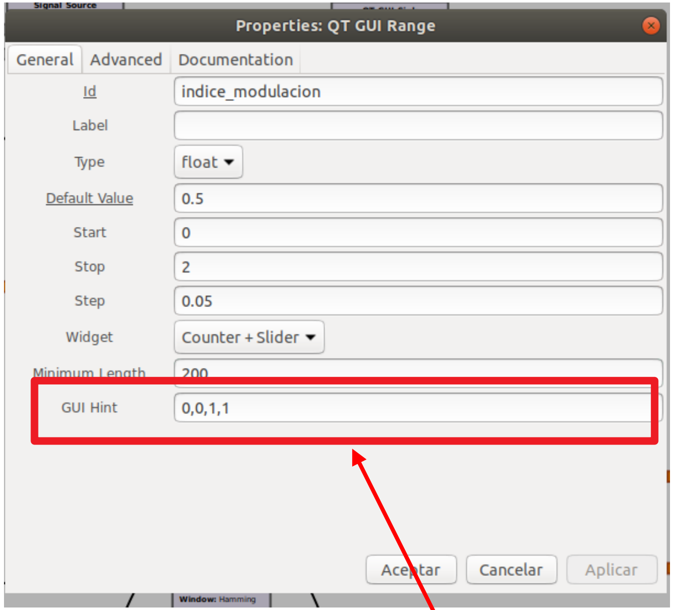

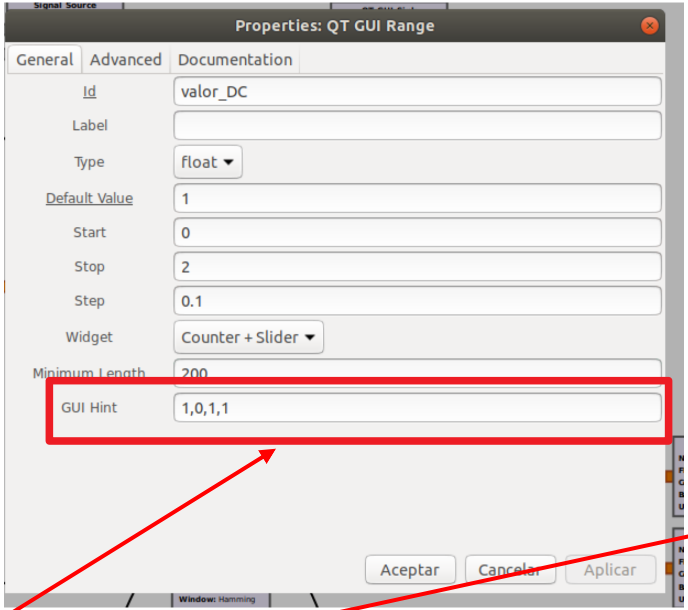

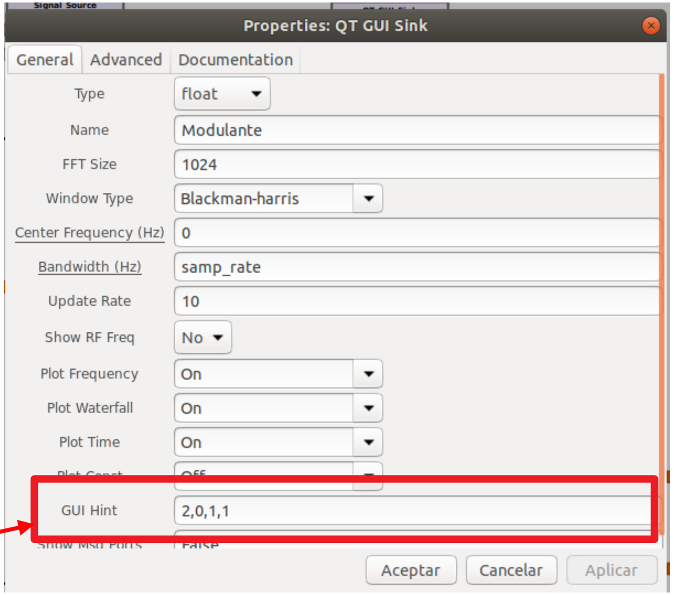

This field allows you to define the position of each QT GUI (scroll bars, graphics, etc) on the screen

Format:

(0,0)	(0,1)	(0,2)	(0,3)
(1,0)	(1,1)	(1,2)	(1,3)
(2,0)	(2,1)	(2,2)	(2,3)

Waveform Selector (0,0,2,1)	Offset Slider (0,1,1,1)
	Frequency Slider (1,1,1,1)
Time Display (2,0,1,1)	Frequency Display (2,1,1,1)

Source: https://wiki.gnuradio.org/index.php/GUI_Hint

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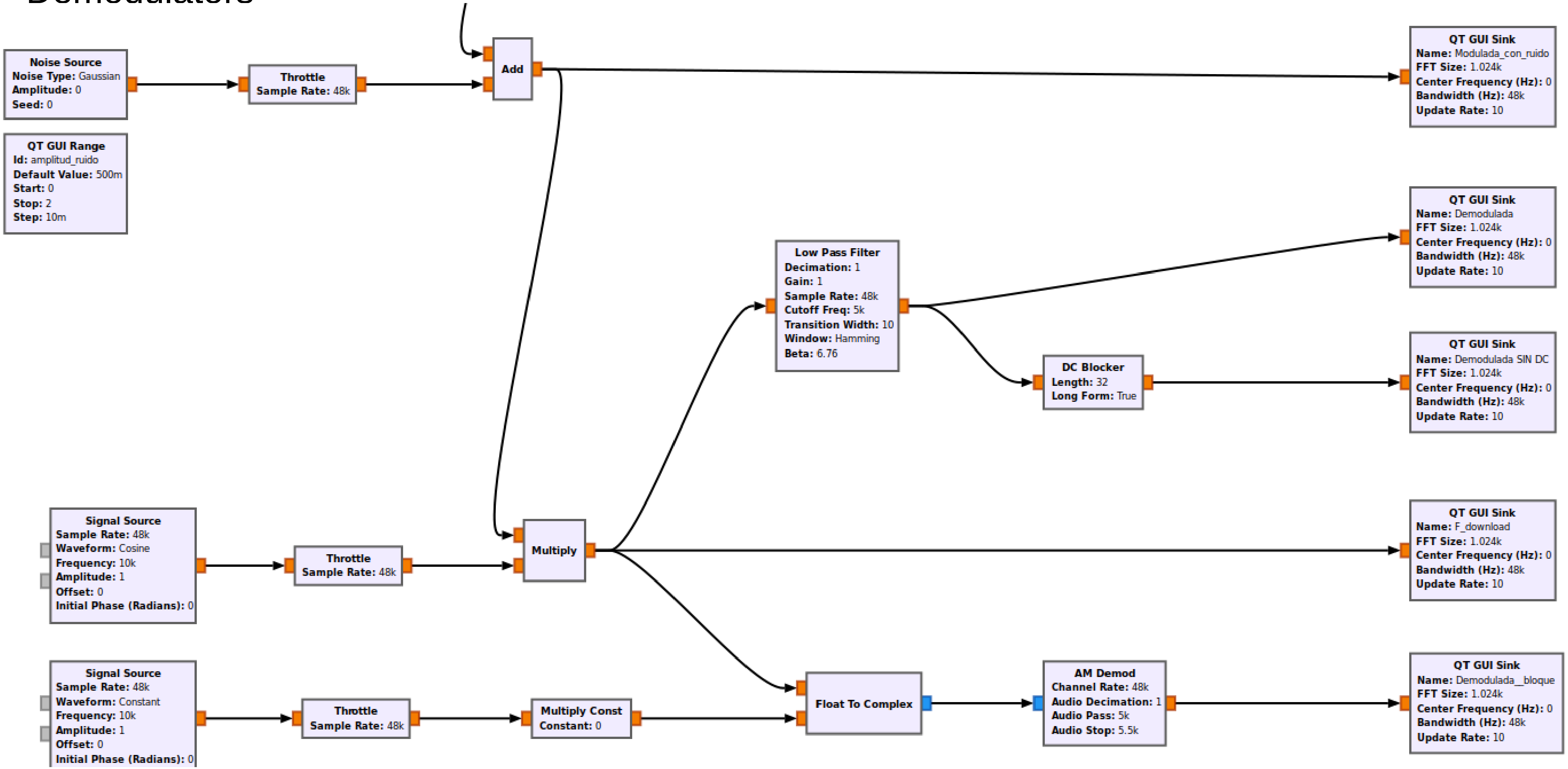
Demodulators



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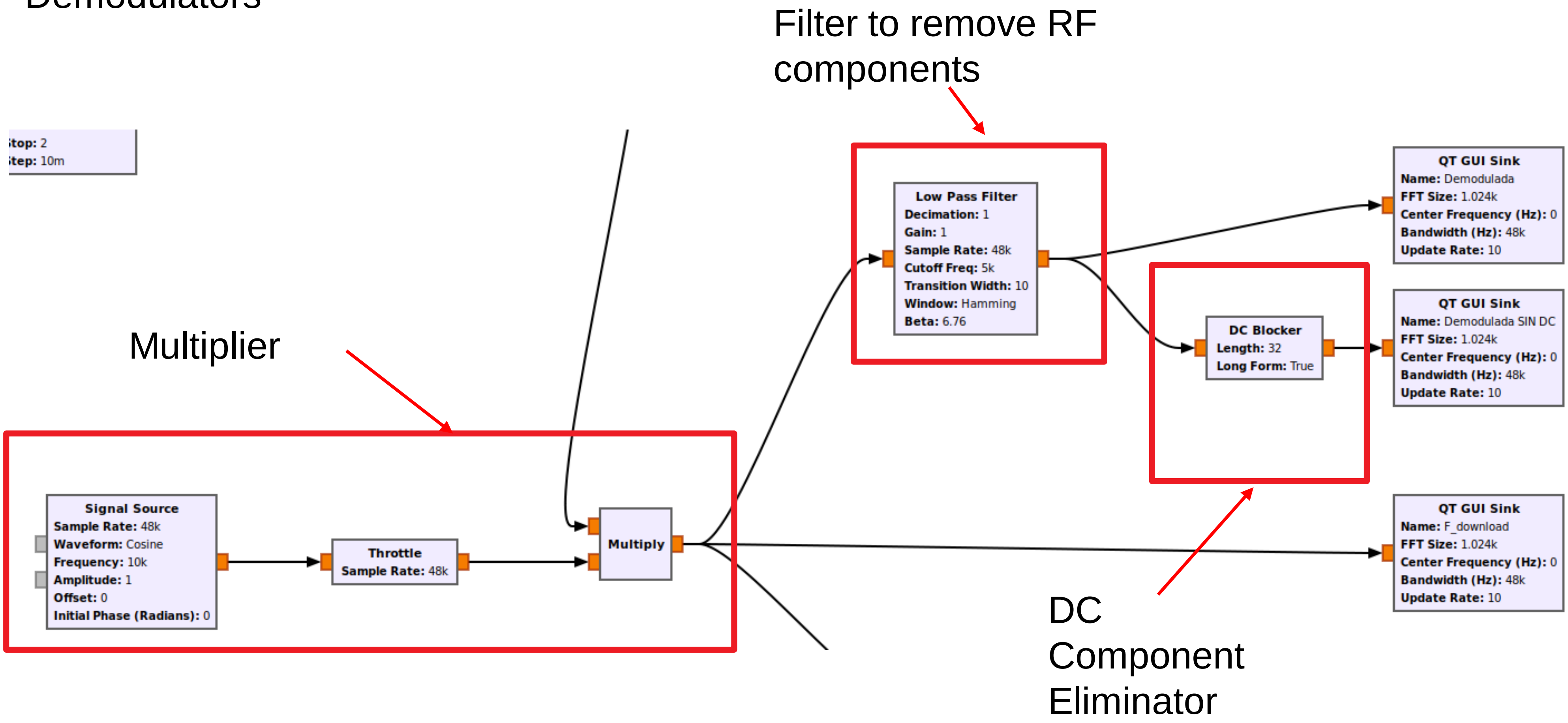
Demodulators



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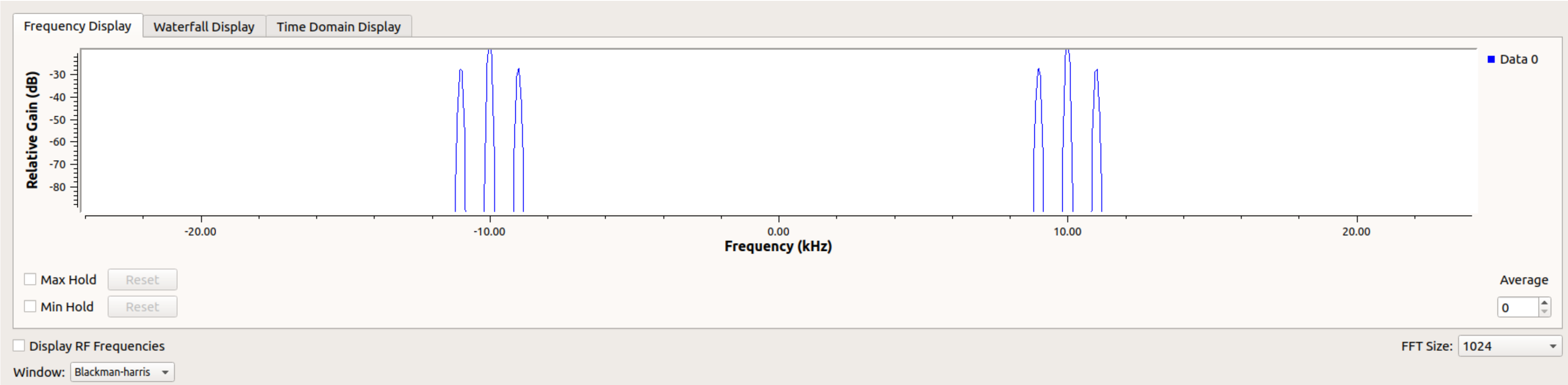
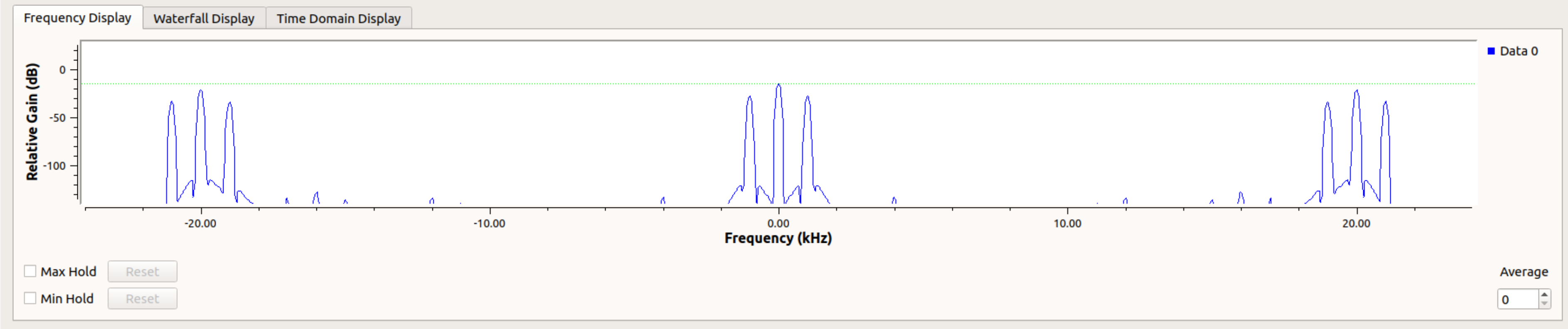
Signal received



Signal multiplier output



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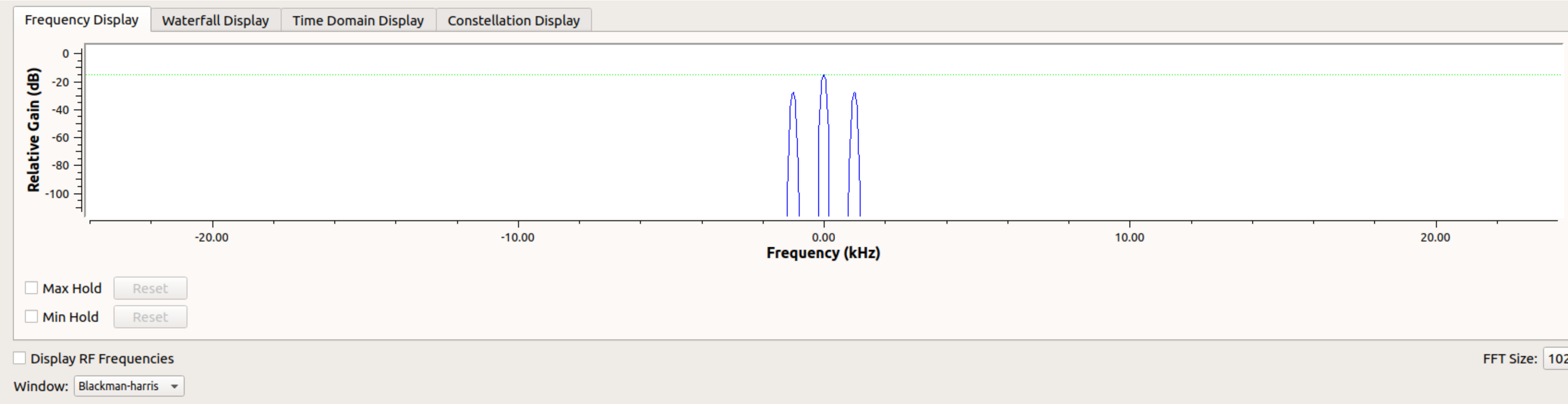
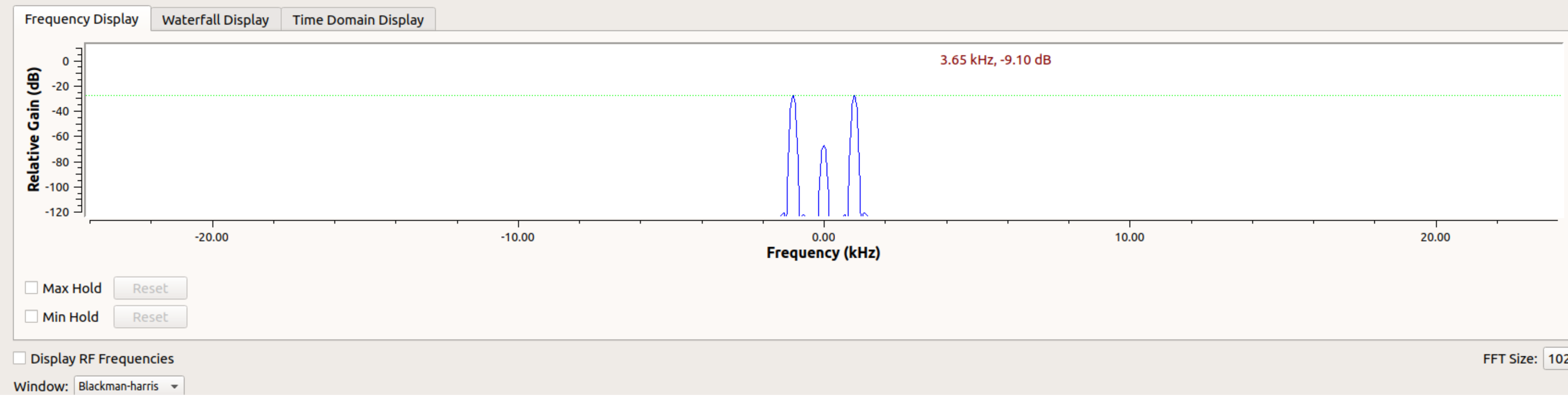
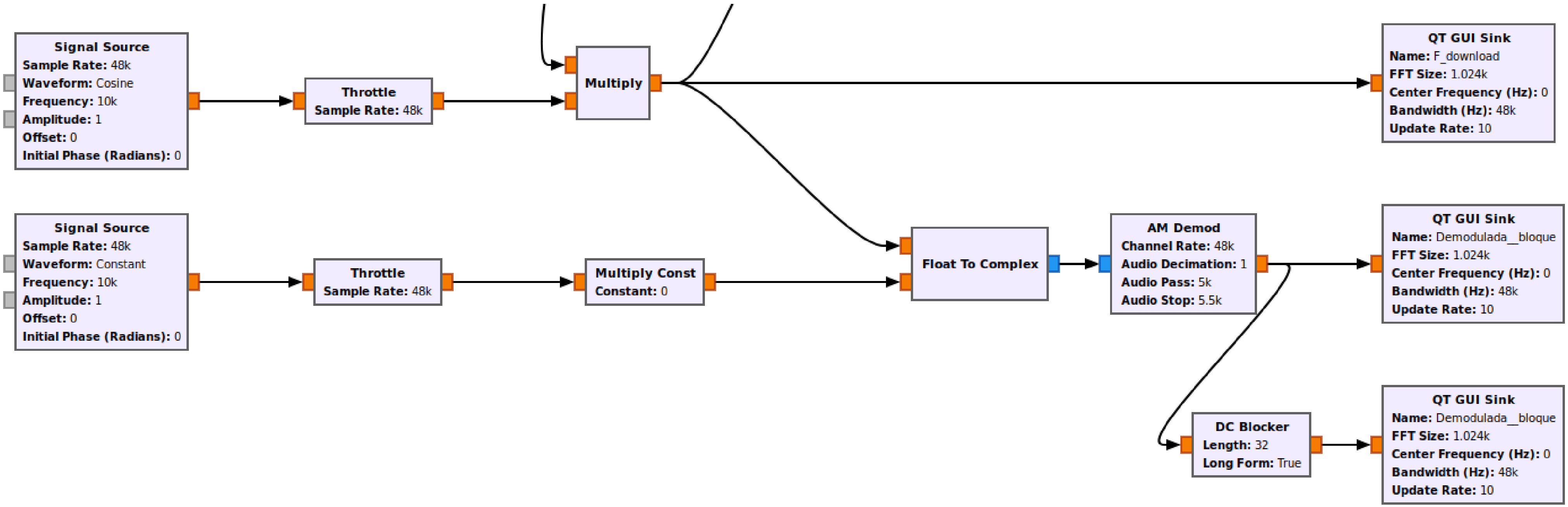
Output filter



After the blocking element of DC. It doesn't totally eliminate it, **but it attenuates 50dB**



If the “AM Demod” modulator from the library is used:



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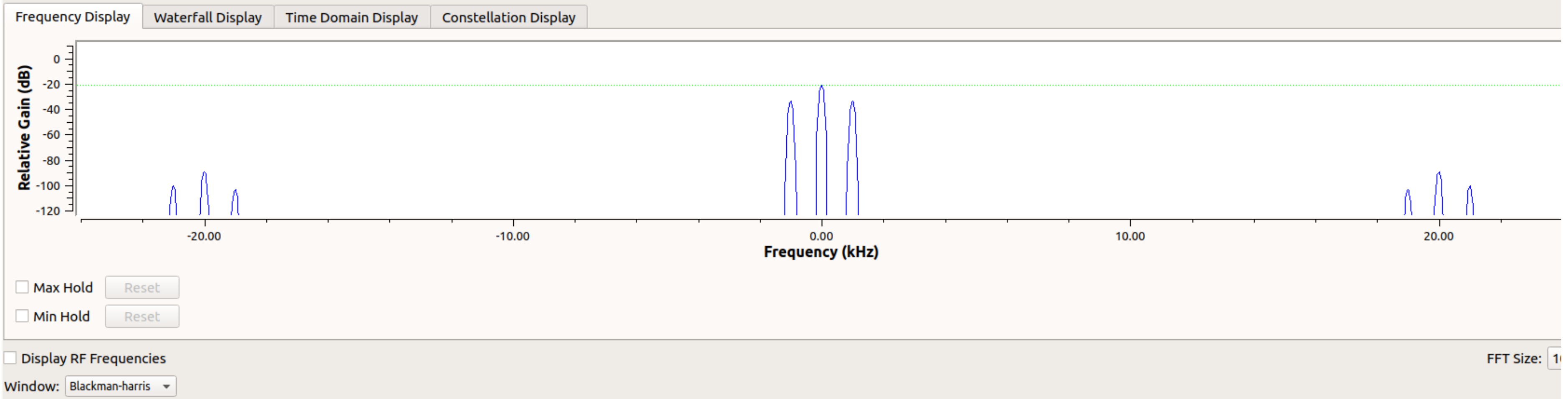
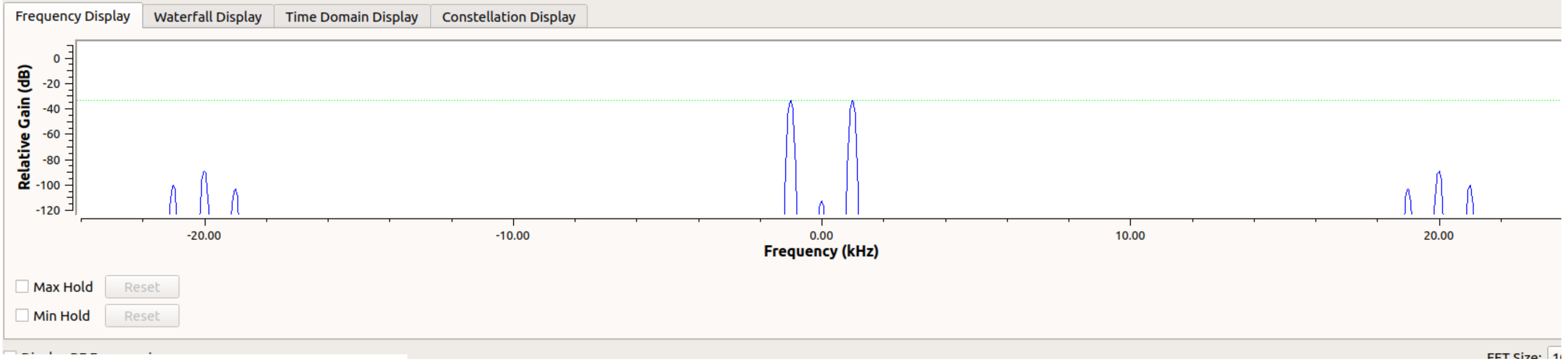
If the “AM Demod” modulator from the library is used:

After the “AM Demod” block. This block does not totally eliminate the RF components.



The DC blocker should also be added, which again heavily attenuates the DC component.



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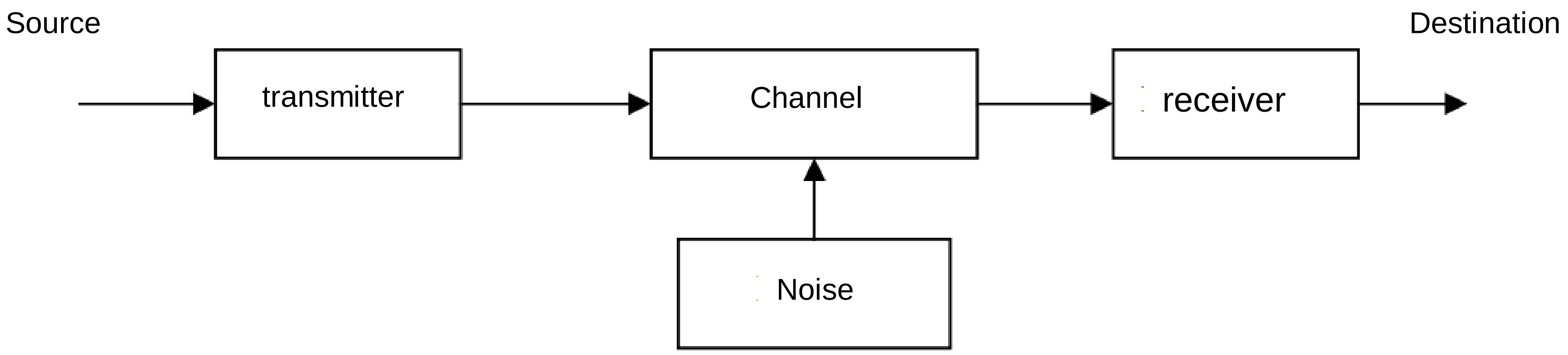
Noise addition:



The classical model assumes noise addition at the receiver input.

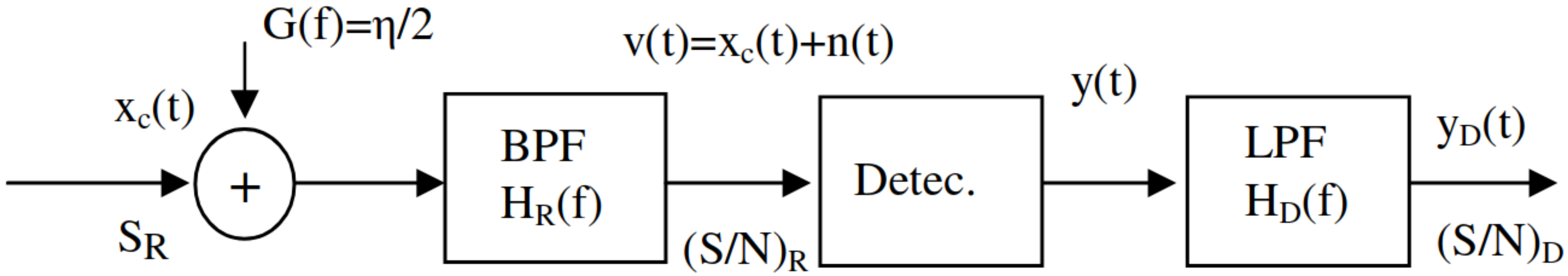


Fig. 8.4 Receiver model of a linear modulation system.

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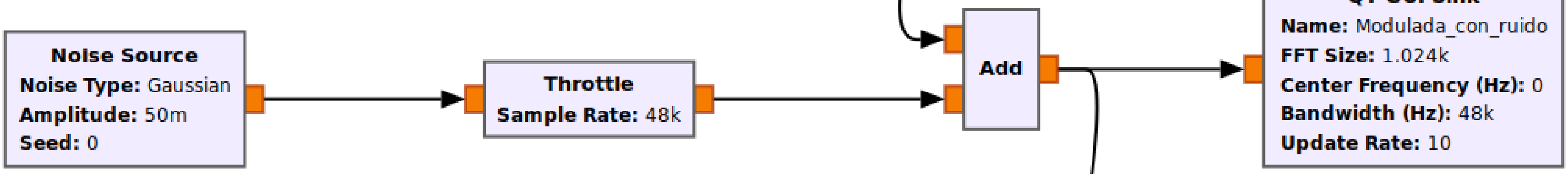
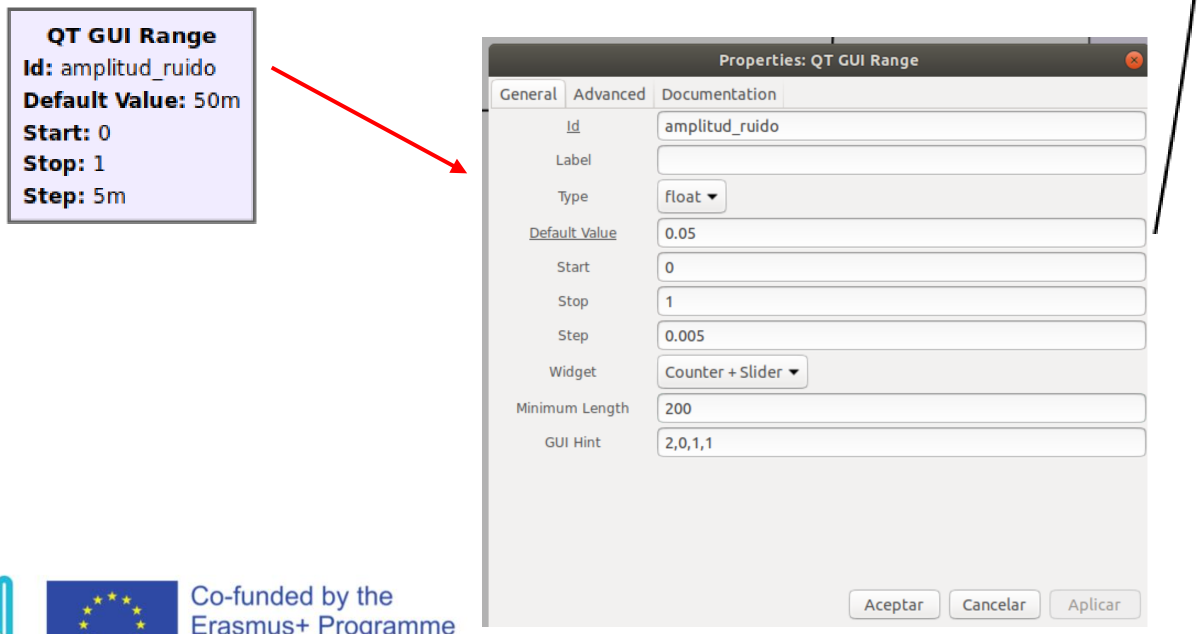
Noise addition:

An AWGN noise generator is added. Its amplitude is adjusted by a variable controllable via a QT GUI slider.





Although synchronous and surround receivers will be used. The amplitude of the noise is adjusted in order to be able to work above and below the threshold.

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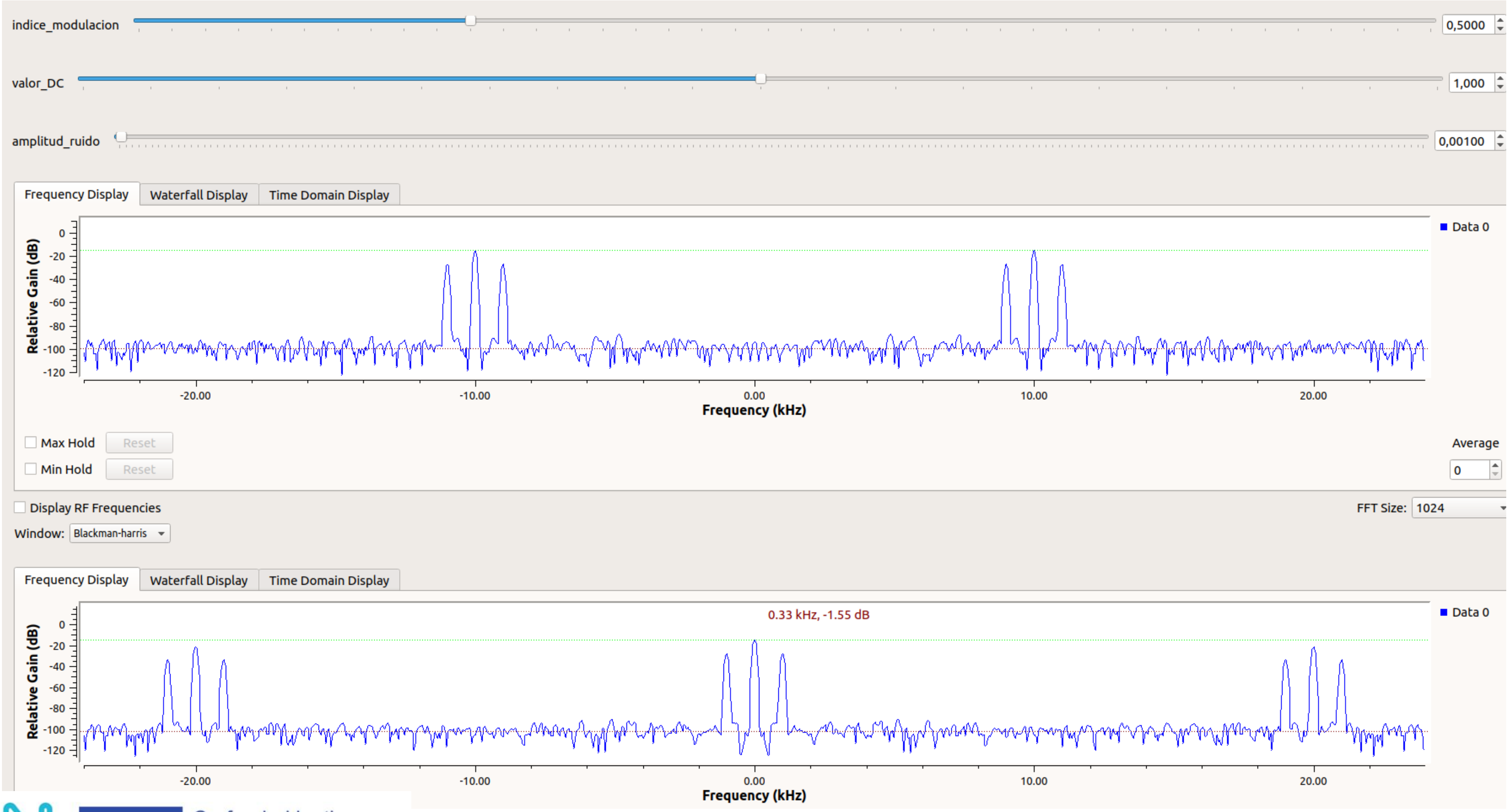
Resulting signals and spectra:

Spectrum Transmitted and received.

Spectrum after multiplier



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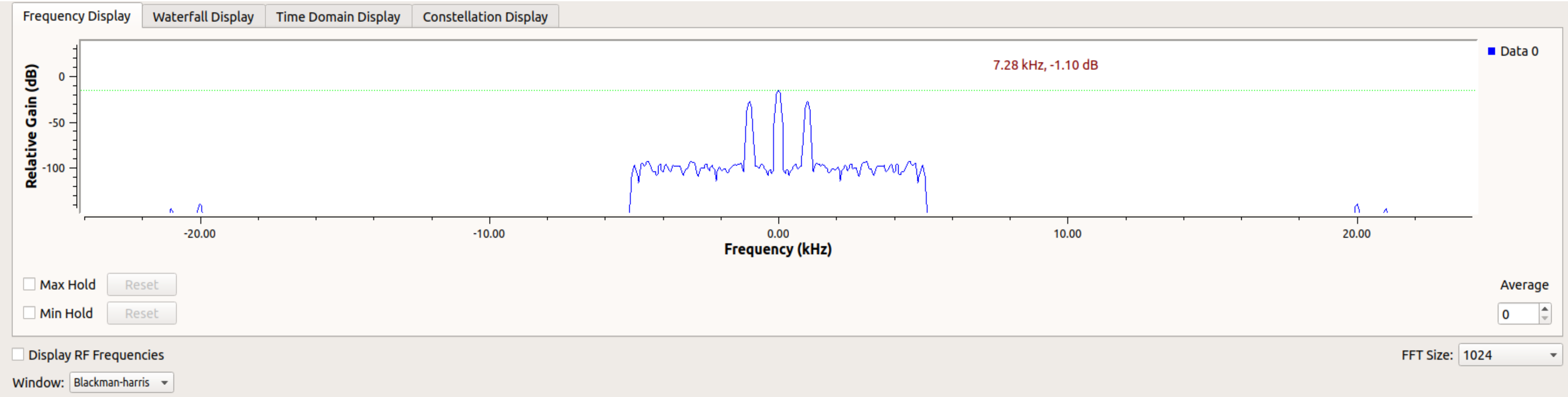
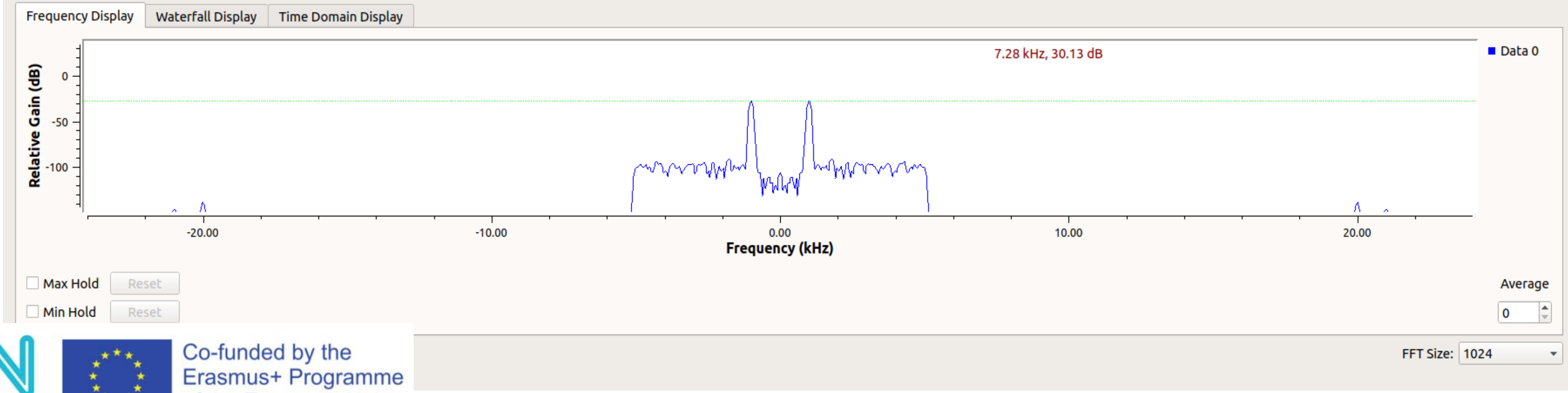
Resulting signals and spectra:

After the filter



After the DC blockade.



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