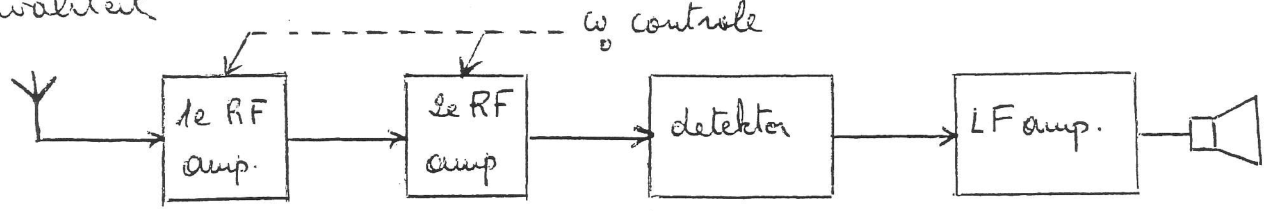
4. Ontvangers

4.1. Ontvangertypes voor AM

4.1.1. De rechtuitontvanger

Dit is de meest eenvoudige ontvanger en tevens deze met mindere kwaliteit



2 of 3 RF-versterkers zijn in kaskade geschakeld, waarbij ze op dezelfde frekwentie afgestemd zijn, resultaat is een hoge versterking (hoge Q-faktor) bij die frekwentie waardoor gevoelige ontvangers kunnen gebouwd worden, doch met het gevaar voor onstabiliteit. Dit HF-signaal wordt via een detektor omgezet in LF waarna het versterkt wordt en uitgeeft op een luidspreker.

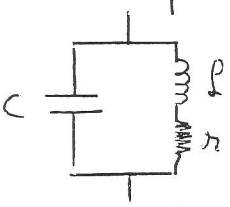
Het grootste nadeel van dit type ontvanger is echter de veranderlijke bandbreedte in funktie van de afstemfrekwentie ω_0.

Inderdaad, stel dat we een AM-ontvanger (500-1500 kHz) wensen te bouwen met een bandbreedte van 10 kHz per zender.

Bij 500 kHz dient een Q-faktor $Q = \frac{f}{\Delta f} = \frac{500}{10} = 50$ aangewend,

terwijl bij 1500 kHz we een andere situatie verkrijgen: bij 1500 kHz is L en Q toegenomen met $\frac{1500}{500} = 3$, zodat theoretisch $Q = 150$ bedraagt, in de praktijk echter zal de serieresistantie van de spoel r echter ook stijgen door grotere verliezen bij hogere frekwenties, zodat $Q = \frac{\omega L}{r}$ misschien 130 bedraagt.

In de praktijk stemt hiermee een bandbreedte van



$$\Delta f = \frac{f}{Q} = \frac{1500}{130} = 11{,}5 \text{ kHz}$$

zodat de kans bestaat dat ook naastliggende zenders ontvangen worden.

Bij de AM-omroepband (500-1500 MHz) is de afwijking nog aanvaardbaar (1,5 kHz op 10 kHz), doch bij HF ontvangers die AM en HF band kombineren zou wanneer we dezelfde bandbreedte van 10 kHz aanhouden bij 30 MHz een Q eisen van

$$\frac{30\,000}{10} = 3000$$

Dit is in de praktijk onmogelijk met RLC-circuits te realiseren. Vandaar dat dit type ontvanger weinig gebruikt wordt tegenwoordig (uitzondering Walkie-Talkies).

4.1.2. Superheterodyne-ontvanger.

Bij de superheterodyne-ontvanger doen we aan frekwentiekonversie, dit betekent dat we alle frekwenties van de te ontvangen band konverteren naar een vaste frekwentie, middenfrekwentie genoemd (engels "Intermediate"-frekwentie of I.F.). Dit middenfrekwent signaal bevat dezelfde modulatie van het oorspronkelijk signaal. Als middenfrekwenties gebruikt men meestal 485 kHz (AM-omroep) en 10,7 MHz (FM-omroep).

De frekwentiekonversie gebeurt in een mengtrap (mixer) met als uitgang de som en verschilfrekwentie. De versterker die op de mixer volgt zorgt dat enkel de verschilfrekwentie versterkt wordt. Het signaal dat aan de menger wordt gelegd is enerzijds f_c en anderzijds f_o van de "lokale oscillator".

Kiest men $f_o > f_c$ dan spreekt men van heterodyne

Kiest men $f_o < f_c$ dan spreekt men van infradyne.

In de praktijk kiest men altijd f_o f_c.

Het verschil $f_o - f_c$ is steeds gelijk aan dezelfde waarde:

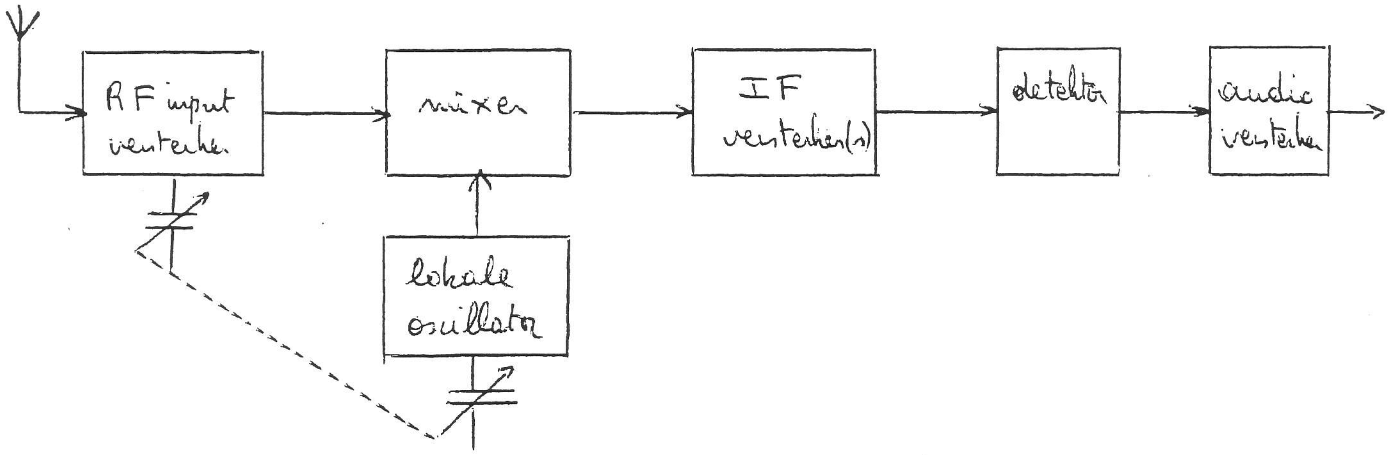
$$f_o - f_c = I.F.$$

Dit betekent dat de afstemming van de zender gebeurt door verandering van de frekwentie van de lokale oscillator.

Door deze afstemknop te verdraaien wijzigt men ook meteen de afstemfrekwentie van de ingangs RF-versterker (dubbel uitgevoerde afstemkondensator). Deze RF-versterker bezit een brede doorlaatband (dus lage Q-faktor) en heeft enkel tot taak ruis en interferentie te vermijden.

De eigenlijke selektiviteit gebeurt in de middenfrekwent versterker(s)

die op 1 frekwentie werk(t)(en) en waarvan de Q-faktor dus hoog kan genomen worden, zonder bandbreedte verandering.
Het blokschema is hieronder aangegeven.



Hierna volgt nu de werking van de individuele blokken voor AM-ontvangers.

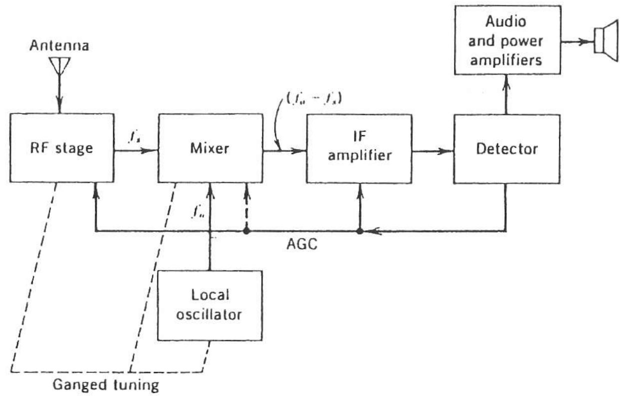


Fig. 7-2 The superheterodyne receiver.

A constant frequency difference is maintained between the local oscillator and the RF circuits, normally through capacitance tuning, in which all the capacitors are *ganged* together and operated in unison by one control knob. The IF amplifier generally uses two or three transformers, each consisting of a pair of mutually coupled tuned circuits. With this large number of double-tuned circuits operating at a constant, specially chosen frequency, the IF amplifier provides most of the gain (and therefore sensitivity) and bandwidth requirements of the receiver. Since the characteristics of the IF amplifier are independent of the frequency to which the receiver is tuned, the selectivity and sensitivity of the superhet are usually fairly uniform throughout its tuning range, and not subject to the variations that beset the TRF receiver. The RF circuits are now used mainly to select the wanted frequency, to reject interference such as the *image frequency* and (especially at high frequencies) to reduce the noise figure of the receiver.

The advantages of the superheterodyne receiver make it the most suitable type for the great majority of radio receiver applications; AM, FM, communications, single-sideband, television and even radar receivers all use it, with only slight modifications in principle. It may be considered as today's standard form of radio receiver, and as such it will now be examined in some detail, section by section.

7-2 AM RECEIVERS

Since the type of receiver is much the same for the various forms of modulation, it has ben found most convenient to explain the principles of a superheterodyne receiver in general while dealing with AM receivers in particular. In this way, a basis is formed with the aid of a simple example of the use of the superheterodyne principle, so that more com-

plex versions can be compared and contrasted with it afterwards; at the same time the overall system is being treated from a practical point of view.

4.1.3. Onderdelen van de Superheterodyne ontvanger

7-2.1 RF Section and Characteristics

A radio receiver always has an RF section, which is a tuned (and tunable) circuit, connected to the antenna terminals. It is there to select the wanted frequency and reject some of the unwanted frequencies. However, such a receiver need not have an RF amplifier following this tuned circuit. If there is an amplifier, its output is fed to the mixer, at whose input another tunable circuit is present. In many instances, however, the tuned circuit connected to the antenna is the actual input circuit of the mixer; the receiver is then said to have no RF amplifier or, more simply, no RF stage.

Reasons for use and functions of RF amplifier The receiver having an RF stage is undoubtedly superior in performance to the receiver without one, all else being equal. On the other hand, there are some instances in which an RF amplifier is uneconomical, i.e., where its inclusion would increase the cost of the receiver significantly but would improve performance only marginally. The best example of this kind of receiver is one which is used for entertainment purposes in a high-signal-strength area, such as the metropolitan area of any large city.

The benefits accruing from the use of an RF amplifier are as follows (reasons 4 to 7 are either more specialized or less important):

1. Greater gain, i.e., better sensitivity
2. Improved image-frequency rejection
3. Improved signal-to-noise ratio
4. Improved rejection of adjacent unwanted signals, i.e., better selectivity
5. Better coupling of the receiver to the antenna (important at VHF and above)
6. Prevention of spurious frequencies from entering the mixer and heterodyning there to produce an interfering frequency equal to the IF from the desired signal
7. Prevention of reradiation of the local oscillator through the antenna of the receiver (rare)

The single-tuned, transformer-coupled type is the amplifier most commonly employed for RF amplification, as illustrated in Fig. 7-3. Both diagrams in the figure are seen to have an RF gain control, which is very rare with domestic receivers but quite common in communications receivers. Whereas the medium-frequency amplifier of Fig. 7-3*a* is quite

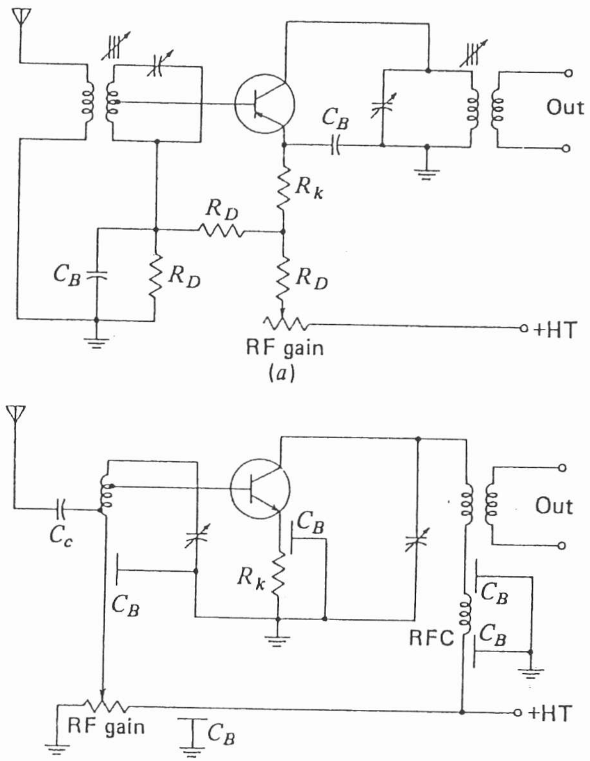


Fig. 7-3 Transistor RF amplifiers. (*a*) Medium frequency; (*b*) VHF.

straightforward, the VHF amplifier of Fig. 7-3*b* contains a number of refinements. Feedthrough capacitors are used as bypass capacitors and, in conjunction with the RF choke, to decouple the output from the HT. Such feedthrough capacitors are almost invariably provided for bypassing at VHF, and often have a value of 1000 pF. In addition, a single-tuned circuit is used at the input, and is coupled to the antenna by means of a trimmer (the latter being manually adjustable for matching to different antennas). Such coupling is used here because of the high frequencies involved. It should also be mentioned that integrated circuits, rather than discrete ones as shown, are used in some receivers. Finally, RF amplifiers have the input and output tuning capacitors ganged to each other and to the one tuning the local oscillator.

Sensitivity The sensitivity of a radio receiver is its ability to amplify weak signals. It is often defined in terms of the voltage that must be applied to the receiver input terminals to give a standard output power, measured at the output terminals. For AM broadcast receivers several

of the relevant quantities have been standardized. Thus 30 percent modulation by a 400-Hz sine wave is used, and the signal is applied to the receiver through a standard coupling network known as a *dummy antenna*. The standard output is 50 mW, and for all types of receivers the loudspeaker is replaced by a load resistance of equal value.

Sensitivity is often expressed in microvolts or in decibels below one volt, and measured at three points along the tuning range when a production receiver is lined up. It is seen that Fig. 7-4 shows the sensitivity curve to vary over the tuning band. At 1000 kHz input frequency, this particular receiver has a sensitivity of 12.7 μV, or −98 dBV (dB below 1 V). Sometimes the above definition is extended, and a manufacturer may quote the sensitivity to be, not merely 12.7 μV for this receiver, but "12.7 μV for a signal-to-noise ratio of 20 dB in the output of the receiver." For professional receivers, there is a tendency to quote the sensitivity in terms of signal power required to produce a minimum acceptable output signal with a minimum acceptable output noise level. The measurements are made under the conditions described. For instance, if the sensitivity of the receiver of Fig. 7-4 at 1000 kHz were to be quoted in this way, we might assume that its input impedance was 50 Ω, and 50 mW happened to be the minimum acceptable value for the (output) signal-to-noise ratio. The minimum input power would thus be $P = E^2/R = (12.7 \times 10^{-6})^2/50 = 3.23 \times 10^{-12} = 3.23$ pW. This is awkward, and is best converted to dB below 1 mW or dBm. Finally, therefore, under the heading of "Sensitivity" in the specifications of a receiver, a manufacturer might quote "a −85-dBm 1-MHz signal, 30 percent modulated with a 400-Hz sine wave will, when applied to the input terminals of this receiver through a dummy antenna, produce an output of at least 50 mW with a signal-to-noise ratio not less than 20 dB in the output."

The most important factors determining the sensitivity of a superheterodyne receiver are the gain of the IF amplifier(s) and that of the RF amplifier, if there is one. It is also obvious from the foregoing that the noise figure plays an important part. Figure 7-4 shows the noise figure plot of a rather good domestic or car radio. Portable and other small receivers used only for the broadcast band might have a sensitivity

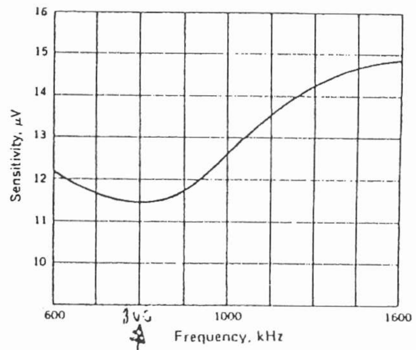


Fig. 7-4 Sensitivity curve for good domestic receiver.

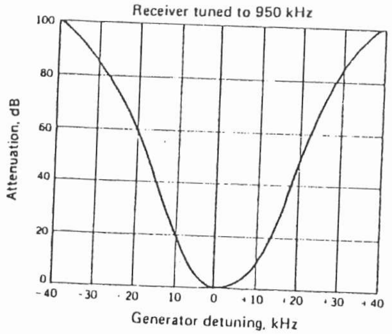


Fig. 7-5 Typical selectivity curve.

in the vicinity of 150 μV, whereas the sensitivity of quality communications receivers may be below 1 μV in the HF band.

Selectivity The selectivity of a receiver is its ability to reject (adjacent) unwanted signals. It is expressed as a curve, such as the one of Fig. 7-5, which virtually shows the attenuation that the receiver offers to signals at frequencies adjacent to the one to which it is tuned. Selectivity is measured at the end of a sensitivity test with conditions the same as for sensitivity, except that now the frequency of the generator is varied to either side of the frequency to which the receiver is tuned. Naturally the output of the receiver falls since the input frequency is now incorrect. Thus the input voltage must be increased until the output is the same as it was originally. The ratio of the voltage required off resonance to the voltage required when the generator is tuned to the receiver's frequency is calculated at a number of points, and then plotted in decibels to give a curve, of which the one in Fig. 7-5 is representative. Looking at the curve, we see that, for example, at 20 kHz below the receiver tuned frequency, an interfering signal would have to be 60 dB greater than the wanted signal to come out with the same amplitude.

Selectivity varies with receiving frequency, and becomes somewhat worse when the receiving frequency is raised. In general, it is determined by the response of the IF section, with the mixer and RF amplifier input circuits playing a small but significant part. It should be noted that it is selectivity that determines the adjacent-channel rejection of a receiver.

Image frequency and its rejection In a standard broadcast receiver (and, in fact, in the vast majority of all receivers made) the local oscillator frequency is made higher than the incoming signal frequency for reasons that will become apparent. It is made equal to the signal frequency plus the intermediate frequency at all times. Thus $f_o = f_s + f_i$, or $f_s = f_o - f_i$, no matter what the signal frequency may be. When f_s and f_o are mixed in the frequency changer, the difference frequency, which is one of the byproducts, is equal to f_i. As such, it is the only one passed and amplified by the IF stage.

If a frequency f_{si} manages to reach the mixer, such that $f_{si} = f_o + f_i$, that is, $f_{si} = f_s + 2f_i$, then this frequency will also produce f_i when mixed with f_o. Unfortunately, this spurious intermediate frequency signal will also be amplified by the IF stage, and will therefore provide interference. This has the effect of two stations being received simultaneously and is naturally undesirable. f_{si} is called the *image frequency*, and is defined as the signal frequency plus twice the intermediate frequency. Reiterating, we have

$$f_{si} = f_s + 2f_i \qquad (7\text{-}1)$$

The rejection of an image frequency by a single-tuned circuit, i.e., the ratio of the gain at the signal frequency to the gain at the image frequency, is given by

$$\alpha = \sqrt{1 + Q^2\rho^2} \qquad (7\text{-}2)$$

where

$$\rho = \frac{f_{si}}{f_s} - \frac{f_s}{f_{si}} \qquad (7\text{-}3)$$

Q = loaded Q of tuned circuit

If the receiver has an RF stage, then there are two tuned circuits, both tuned to f_s; the rejection of each will be calculated by the same formula, and the total rejection will be the product of the two. Whatever applies to gain calculations applies also to those involving rejection.

Image rejection depends on the front-end selectivity of the receiver and *must be achieved before the IF stage.* Once the spurious frequency enters the first IF amplifier, it becomes impossible to remove it from the wanted signal. It can be seen that if f_{si}/f_s is large, as it is in the broadcast band, the use of an RF stage is not essential for good image-frequency rejection, but it does become necessary in the short-wave range and beyond.

Example 7-1 In a broadcast superheterodyne receiver having no RF amplifier, the loaded Q of the antenna coupling circuit (at the input to the mixer) is 100. If the intermediate frequency is 455 kHz, calculate (*a*) the image frequency and its rejection ratio at 1000 kHz and (*b*) the image frequency and its rejection ratio at 25 MHz.

(*a*) $f_{si} = 1000 + 2 \times 455 = 1910$ kHz

$$\rho = \frac{1910}{1000} - \frac{1000}{1910} = 1.910 - 0.524 = 1.386$$

$$\alpha = \sqrt{1 + 100^2 \times 1.386^2} = \sqrt{1 + 138.6^2} = 138.6$$

This is 42 dB, and is considered adequate for domestic receivers in the MF band.

(*b*) $f_{si} = 25 + 2 \times 0.455 = 25.91$ MHz

$$\rho' = \frac{25.91}{25} - \frac{25}{25.91} = 1.0364 - 0.9649 = 0.0715$$

$$\alpha = \sqrt{1 + 100^2 \times 0.0715^2} = \sqrt{1 + 7.15^2} = 7.22$$

It is apparent that this rejection is insufficient for a practical receiver in the HF band.

Example 7-1 shows, as it was meant to, that although image rejection need not be a problem for a broadcast receiver without an RF stage, special precautions must be taken at HF. This will be seen in Sec. 7-3, but two possibilities can be explored now, in Example 7-2.

Example 7-2 In order to make the image frequency rejection of the receiver of Example 7-1 as good at 25 MHz as it was at 1000 kHz, calculate (*a*) the loaded Q which an RF amplifier for this receiver would have to have and (*b*) the new intermediate frequency that would be needed (if there is to be no RF amplifier).

(*a*) Since the mixer already has a rejection of 7.22, the image rejection of the RF stage will have to be

$$\alpha' = \frac{138.6}{7.22} = 19.2 = \sqrt{1 + Q'^2 \times 0.0715^2}$$

$$Q'^2 = \frac{19.2^2 - 1}{0.0715}$$

$$Q' = \frac{\sqrt{367.6}}{0.0715} = 268$$

Understandably, of course, a well-designed receiver would have the same Q for both tuned circuits. Here this works out to 164 each, that being the geometric mean of 100 and 268.

(*b*) If the rejection is to be the same as initially, through a change in the intermediate frequency, it is apparent that ρ will have to be the same as in Example 7-1 *a*, since the Q is also the same. Thus

$$\frac{f'_{si}}{f'_s} - \frac{f'_s}{f'_{si}} = 138.6 = \frac{1910}{1000} - \frac{1000}{1910}$$

$$\frac{f'_s}{f'_{si}} = \frac{1910}{1000} = 1.91$$

$$\frac{25 + 2f'_i}{25} = 1.91$$

$$25 + 2f'_i = 1.91 \times 25$$

$$f'_i = \frac{1.91 \times 25 - 25}{2} = \frac{0.91 \times 25}{2} = 11.4 \text{ MHz}$$

Double spotting This is a well-known phenomenon, which manifests itself by the picking up of the same short-wave station at two nearby

points on the receiver dial. It is caused by poor front-end selectivity, i.e., inadequate image-frequency rejection. That is to say, the front end of the receiver does not select different adjacent signals very well, but, fortunately, the IF stage takes care of eliminating almost all of them. This being the case, it is obvious that the precise tuning of the local oscillator is what determines which signal will be amplified by the IF stage. Within broad limits, the setting of the tuned circuit at the input of the mixer is far less important (it being assumed that there is no RF amplifier in a receiver which badly suffers from double spotting). Consider such a receiver at HF, having an IF of 455 kHz. If there is a strong station at (say) 14.7 MHz, the receiver will naturally pick it up—note that, when it does, the local oscillator frequency will be 15.155 MHz. However, the receiver will also pick up this strong station when it (the receiver) is tuned to 13.790 MHz. When the receiver is tuned to the second frequency, its local oscillator will be adjusted to 14.245 MHz. Since this is exactly 455 kHz below the frequency of the strong station, the two signals will produce 455 kHz when they are mixed, and of course the IF amplifier will not reject this signal. If there had been an RF amplifier, the 14.7-MHz signal might have been rejected before reaching the mixer, but without an RF amplifier this receiver cannot adequately reject 14.7 MHz when it is tuned to 13.79 MHz.

Double spotting is harmful to the extent that a weak station may be masked by the reception of a nearby strong station at the spurious point on the dial. As a matter of fact, double spotting may be used to calculate the intermediate frequency of an unknown receiver, since the spurious point on the dial is precisely $2f_i$ *below* the correct frequency.

As expected, an improvement in image-frequency rejection will produce a corresponding reduction in double spotting.

7-2.2 Frequency Changing and Tracking

Generally speaking, a frequency changer[1] is a nonlinear resistance having two sets of input terminals and one set of output terminals. The signal from the antenna or from the preceding RF amplifier is fed to one set of input terminals, while the output of the local oscillator is fed to the other set. As was shown in Eq. (6-8), such a nonlinear resistance will have several frequencies present in its output, including the difference between the two input frequencies—in modulation work this was called the lower sideband. The difference frequency here is the intermediate frequency, and is the one to which the output circuit of the mixer is tuned.

The most common types of mixers are the bipolar transistor, FET and integrated circuit. All three are generally self-excited, so that the

[1] More commonly called a *mixer*, sometimes a *converter*, and, in the early days of radio, *the first detector*.

device acts as both oscillator and mixer. When tubes were common, the pentagrid and triode-hexode were made specially for self-excited mixer duty. At UHF and above, crystal (i.e., silicon) diodes have been used as mixers since before World War II, because of their low noise figures. These and other diodes, with even lower noise figures, are still so used, as will be seen in the microwave chapters. Naturally, diode mixers are separately excited.

Conversion transconductance It will be recalled that the coefficient of nonlinearity of most nonlinear resistances is rather low, so that the IF output of the mixer will be very low indeed unless some preventive steps are taken. The usual step is to make the local oscillator voltage quite large, 1 V rms or more to a mixer whose signal input voltage might be 100 μV or less. That this has the desired effect is shown by term (V) of Eq. (6-8). It is then said that the local oscillator *varies the bias* on the mixer from zero to cutoff, thus varying the transconductance in a nonlinear manner. The mixer amplifies the signal with this varying g_m, and an IF output results.

Like any other amplifying device, a mixer has a transconductance. However, the situation here is a little more complicated, since the output frequency is different from the input frequency. *Conversion transconductance* is defined as

$$g_c = \frac{\Delta i_p \text{ (at the intermediate frequency)}}{\Delta e_g \text{ (at the signal frequency)}} \qquad (7\text{-}4)$$

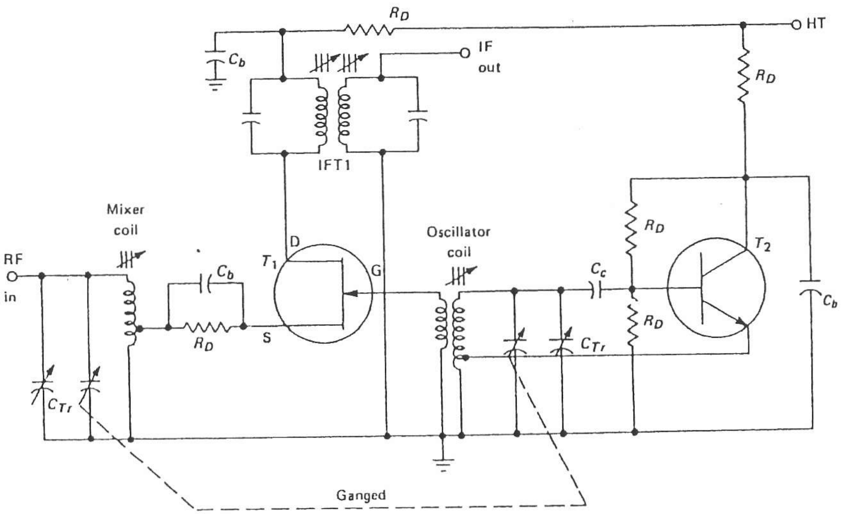


Fig. 7-6 Separately excited FET mixer.

The conversion transconductance of a transistor mixer is of the order of 6 mS, which is decidedly lower than the g_m of the same transistor used as an amplifier. Since g_c depends on the size of the local oscillator voltage, the above value refers to optimum conditions.

Separately excited mixer In this circuit, which is shown in Fig. 7-6, one device acts as a mixer while the other supplies the necessary oscillations. In this case, T_1, the FET, is the mixer, to whose gate is fed the output of T_2, the bipolar transistor Hartley oscillator. A FET is well suited for mixer duty, because of the square-law characteristic of its drain current. Note the ganging together of the tuning capacitors across the mixer and oscillator coils, and that each in practice has a trimmer (C_{Tr}) across it for fine adjustment by the manufacturer. Note further that the output is taken via a double-tuned transformer (the first IF transformer) in the drain of the mixer, and fed to the IF amplifier. The arrangement as shown is most common at higher frequencies, whereas in domestic receivers a self-excited mixer is more likely to be encountered.

Self-excited transistor mixer[1] The transistor circuit of Fig. 7-7 is best considered at each frequency in turn. First, however, the significance of the L_5–L_3 arrangement must be explained; it is necessary that the tuned circuit L_3–C_G be placed between collector and ground, but only for ac purposes. Furthermore, the construction of a ganged capacitor (C_G is one of its sections) is such that in all the various sections the rotating plates are connected to one another via the rotor shaft. To avoid difficulties, the rotor of the gang is grounded. Thus one end of C_G must naturally go to

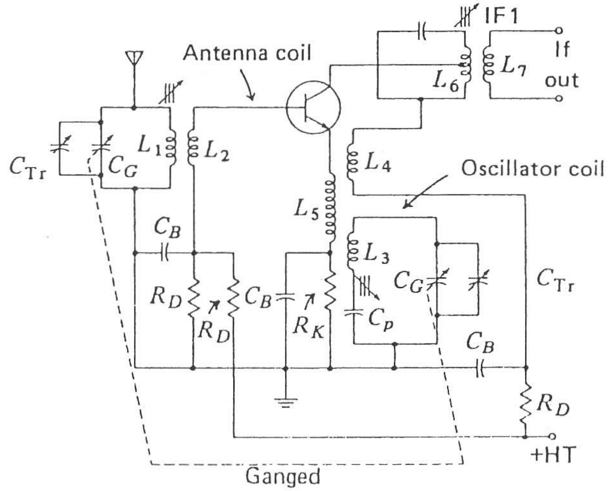


Fig. 7-7 Self-excited bipolar transistor mixer.

ground, and yet there must be a continuous path for direct current from HT to collector. One of the solutions to this problem would be the use of an RF choke instead of L_4, and the connection of a coupling capacitor from the bottom of L_6 to the top of L_3, but the arrangement as shown is equally effective and happens to be simpler and cheaper. It is merely inductive coupling instead of a coupling capacitor, and an extra transformer winding instead of an RF choke.

Now, at the signal frequency, the collector and emitter tuned circuits may be considered as being effectively short-circuited so that (at the RF) we have an amplifier with an input tuned circuit and an output that is indeterminate. At the IF, on the other hand, the base and emitter circuits are the ones which may be considered short-circuited. Thus, at the IF, we have an amplifier whose input comes from an indeterminate source, and whose output is tuned to the IF. Both these "amplifiers" are common-emitter amplifiers.

At the local oscillator frequency, the RF and IF tuned circuits may both be considered as though they were short-circuited, so that the equivalent circuit of Fig. 7-8 results (at f_o only). This is seen to be a tuned-collector Armstrong oscillator of the common-base variety.

We have considered each function of the frequency changer individually, but the circuit performs them all simultaneously, of course. Thus, the circuit oscillates, the transconductance of the transistor is varied in a nonlinear manner at the local oscillator rate, and this variable g_m is used by the transistor to amplify the incoming RF signal. Hence heterodyning occurs, with the resulting production of the required intermediate frequency.

Superheterodyne tracking The superheterodyne receiver (or any receiver for that matter) has a number of tunable circuits which must all be tuned correctly if any given station is to be received. For obvious reasons, the various tuned circuits are coupled mechanically so that only one tuning control and dial are required. In turn, this means that no matter what the received frequency, the RF and mixer input tuned circuits must be tuned to it. The local oscillator must simultaneously be tuned to a frequency precisely higher than this, by the intermediate frequency. Any errors that exist in this frequency difference will result in an incorrect

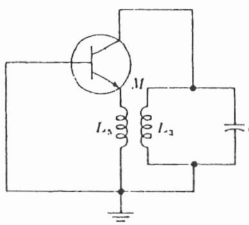


Fig. 7-8 Mixer equivalent at f_o.

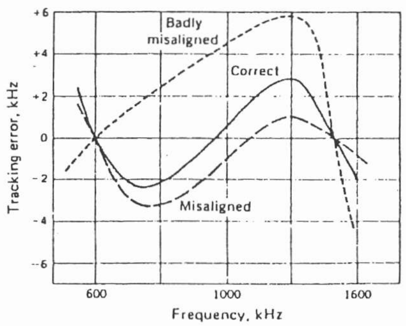


Fig. 7-9 Tracking curves.

frequency being fed to the IF amplifier, and must naturally be avoided. Such errors as exist are called *tracking errors*, and they result in stations appearing away from their correct position on the dial.

Keeping a constant frequency difference between the local oscillator and the front-end circuits is possible neither in theory nor in practice; thus some tracking errors must always occur. The best that can be accomplished is a difference frequency that is equal to the IF at two preselected points on the dial, along with some errors at all other points. However, if a coil is placed in series with the local oscillator ganged capacitor, or, more commonly, a capacitor in series with the local oscillator coil, then *three-point tracking* results and has the appearance of the solid curve of Fig. 7-9. The capacitor in question is called a *padding capacitor* or a *padder* and is shown (labeled C_p) in Figs. 7-6 and 7-7. The wanted result has been obtained because the variation of the local oscillator coil reactance with frequency has been altered. The three frequencies of correct tracking may be chosen in the design of the receiver and are often as shown in Fig. 7-9, that is, just above the bottom end of the band (600 kHz), somewhat below the top end (1500 kHz), and at the geometric mean of the two (950 kHz).

It is entirely possible to keep maximum tracking error below 3 kHz, as shown; a value as low as that is generally considered negligible. However, since the padder has a fixed value, it provides correct three-point tracking only if the adjustable local oscillator coil has been preadjusted, i.e., *aligned*, to the correct value. If this has not been done, then incorrect three-point tracking results, or the center point may disappear completely, as shown in Fig. 7-9.

Local oscillator In receivers operating up to the limit of short-wave broadcasting, that is, 36 MHz, the most common types of local oscillators are the Armstrong and the Hartley. The Colpitts, Clapp, or Ultra Audion oscillators are used at VHF and above, with the Hartley also having some use if frequencies do not exceed about 120 MHz. Note that all these oscillators are *LC* and that each employs only one tuned circuit to determine its frequency. Where, for some reason, the frequency stability of the local oscillator must be particularly high, AFC (see Secs. 5-3.3 and 7-3.2) or a frequency synthesizer (see Secs. 3-5 and 7-5.2) may be used. Ordinary local oscillator circuits are shown in Figs. 7-6 and 7-7.

The frequency range of a broadcast receiver local oscillator is calculated on the basis of a signal frequency range from 540 to 1650 kHz, and an intermediate frequency which is (very often) 455 kHz. For the usual case of local oscillator frequency above signal frequency, this range is 995 to 2105 kHz, giving a ratio of maximum to minimum frequencies of 2.2:1. If the local oscillator had been designed to be below signal frequency, the range would have been 85 to 1195 kHz, and the ratio would have been 14:1. The normal tunable capacitor has a capacitance ratio of approximately 10:1, giving a frequency ratio of 3.2:1. Hence the 2.2:1 ratio required of the local oscillator operating above signal frequency is well within range, whereas the other system has a frequency range that cannot be covered in one sweep. This is why the local oscillator frequency is always made higher than the signal frequency in receivers with variable-frequency oscillators.

It may be shown that tracking difficulties would disappear if the frequency ratio (instead of the frequency difference) were made constant. Now, in the usual system, the ratio of local oscillator frequency to signal frequency is $995/540 = 1.84$ at the bottom of the broadcast band, and $2105/1650 = 1.28$ at the top of the band. In a local-oscillator-below-signal-frequency system, these ratios would be 6.35 and 1.38, respectively. This is a much greater variation in frequency ratio, and would result in far more troublesome tracking problems.

7-2.3 Intermediate Frequencies and IF Amplifiers

Choice of frequency The intermediate frequency of a receiving system is usually a compromise, since there are reasons why it should be neither low nor high, nor between the two. The following are the major factors influencing the choice of the intermediate frequency in any particular system:

1. If the intermediate frequency is too high, poor selectivity and poor adjacent-channel rejection result.
2. A high value of intermediate frequency increases tracking difficulties.
3. As the intermediate frequency is lowered, image-frequency rejection becomes poorer. Equations (7-1), (7-2) and (7-3) showed that rejection is improved as the ratio of image frequency to signal frequency is increased, and this, naturally, requires a high intermediate frequency. Extrapolating, it is seen that image-frequency rejection becomes worse as signal frequency is raised, as was shown by Example 7-1*a* and *b*.
4. A very low intermediate frequency makes the selectivity too sharp, cutting off the sidebands. This problem arises because the Q must be low when the IF is low, and hence the gain per stage is low. Thus a designer is more likely to raise the Q than to increase the number of IF amplifiers.
5. If the IF is very low, the frequency stability of the local oscillator must be made correspondingly higher because any frequency drift is now a larger proportion of the low IF than of a high IF.
6. The intermediate frequency must not fall within the tuning range of the receiver, or else instability will occur and heterodyne whistles will be heard, making it impossible to tune to the frequency band immediately adjacent to the intermediate frequency.

Frequencies used As a result of many years' experience, the foregoing requirements have been translated into specific frequencies, whose use is fairly well standardized throughout the world (but by no means compulsory). These are as follows:

1. Standard broadcast AM receivers [tuning to 540 to 1650 kHz, perhaps 6 to 18 MHz, and possibly even the European long-wave band (150 to 350 kHz)] use an IF within the 438- to 465-kHz range, with 455 kHz the most popular frequency and becoming even more so.
2. AM, SSB and other receivers employed for short-wave or VHF reception have a first IF often in the range from about 1.6 to 2.3 MHz. (Such receivers have two or more different intermediate frequencies. See Sec. 7-3.1.)
3. FM receivers using the standard 88- to 108-MHz band have an IF which is almost always 10.7 MHz.
4. Television receivers in the VHF band (54 to 223 MHz) and in the UHF band (470 to 940 MHz) use an IF between 26 and 46 MHz, with about 36 and 46 MHz the two most popular values.
5. Microwave and radar receivers, operating on frequencies in the 1- to 10-GHz range, use intermediate frequencies depending on the application, with 30, 60 and 70 MHz among the most popular.

By and large, services covering a wide frequency range have IFs somewhat below the lowest receiving frequency, whereas other services, especially fixed-frequency microwave ones, may use intermediate frequencies as much as 40 times lower than the receiving frequency.

IF amplifiers The IF amplifier is a fixed-frequency amplifier, with the

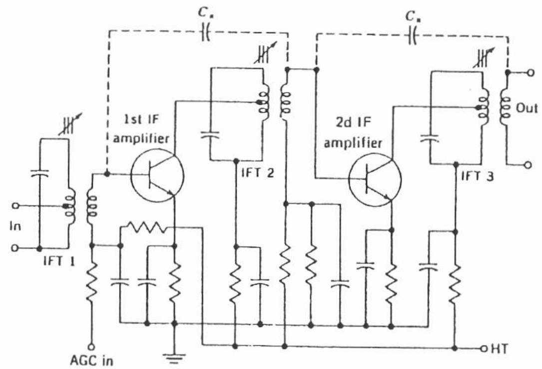


Fig. 7-10 Two-stage IF amplifier.

very important function of rejecting adjacent unwanted frequencies. It should thus have a frequency response with steep skirts. When the desire for a flat-topped response is added, the resulting recipe is for a double-tuned or stagger-tuned amplifier. Whereas FET and integrated circuit IF amplifiers generally are (and vacuum-tube ones always were) double tuned at the input and at the output, bipolar transistor amplifiers often are single tuned. A typical bipolar IF amplifier for a domestic receiver is shown in Fig. 7-10. It is seen to be a two-stage amplifier, with all IF transformers single tuned. This departure from a single-stage, double-tuned amplifier is for the sake of extra gain, and hence receiver sensitivity.

Although a double-tuned circuit rejects adjacent frequencies far better than a single-tuned circuit, bipolar transistor amplifiers, on the whole, use single-tuned circuits for interstage coupling. The reason is simply that greater gain may be achieved in this way because of the need for tapping coils in tuned circuits. This tapping may be required to obtain maximum power transfer and a reduction of the damping of the circuit involved. It will be recalled that the bandwidth of a tuned circuit depends on its loaded Q, which depends on the unloaded Q and the external damping resistance. Since transistor impedances may be low, tapping is employed, together with somewhat lower inductances than would have been used with tube circuits. If a double-tuned transformer is used, both sides of it might have to be tapped, rather than just one side as with a single-tuned transformer. Thus a reduction in voltage would be applied to each transistor electrode, and hence a general reduction in gain. Note also that neutralization may have to be used in the transistor IF amplifier, depending on the frequency and the type of transistor employed.

When double tuning is used, the coefficient of coupling varies from 0.8 times critical to critical; overcoupling is not normally used with-

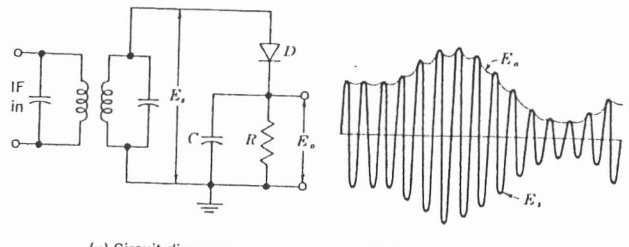


Fig. 7-11 Simple diode detector.

out a special reason. Finally, the IF transformers are often all made identical so that they are interchangeable.

7-2.4 Detection and Automatic Gain Control (AGC)

Operation of diode detector The diode is by far the most common device used for demodulation (or detection), and its operation will now be considered in detail. On the circuit of Fig. 7-11*a*, C is a small capacitance and R is a large resistance; the parallel combination of R and C is the load resistance across which the rectified output voltage E_o is developed. At each positive peak of the RF cycle, C charges up to a potential almost equal to the peak signal voltage E_s. The difference is due to the diode drop, since the forward resistance of the diode is small (but not zero). Between peaks a little of the charge in C decays through R, to be replenished at the next positive peak. The result is the voltage E_o, which reproduces the modulating voltage accurately, except for the small amount of RF ripple. Note that the time constant of the RC combination must be slow enough to keep the RF ripple as small as possible, but sufficiently fast for the detector circuit to follow the fastest modulation variations.

This simple diode detector has the disadvantages that E_o, in addition to being proportional to the modulating voltage, also has a dc component, which represents the average envelope amplitude (i.e., carrier strength), and a small RF ripple. However, the unwanted components are removed in a practical detector, leaving only the intelligence and some second harmonic of the modulating signal.

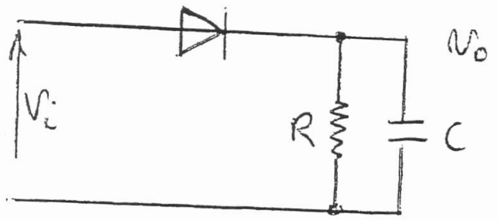
Practical diode detector A number of additions have been made to the simple detector, and its practical version is shown in Fig. 7-12. The circuit operates in the following manner. The diode has been reversed, so that now the negative envelope is demodulated. This has no effect on detection, but it does ensure that a negative AGC voltage will be available, as

4.1.4. Demodulators

De demodulator is de blok die de hoogfrekwente drager waarop het L.F.-signaal gemoduleerd werd (AM, FM, PM) afzondert, zodat enkel de L.F.-informatie overblijft.
Iedere ontvanger zal dus een demodulator bevatten.
Vooraleer men het antennesignaal aansluit op de modulator zal men echter het signaal nog extra versterken, zodat het kompatibel is, kwa niveau, en soms kwa frekwentie, met de demodulatieblok. (zie nr 1.2)

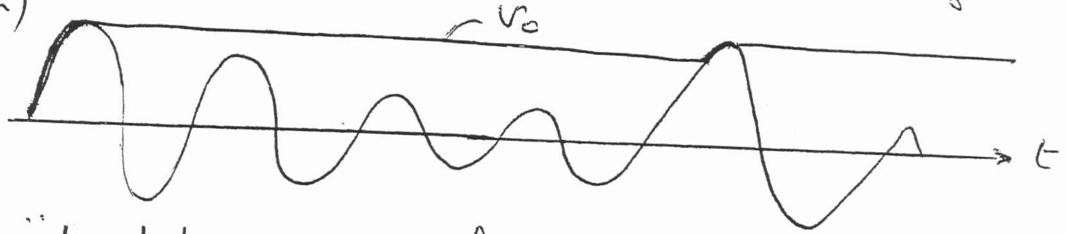
4.1.4.1. Gewone AM-demodulators

1. 1.4.1.1. omhullende-detektor (seriedetektor)



Door een gepaste keuze van R.C. t.o.v. ω_i en ω_C (respektievelijk LF. en HF signaal), volgt v_o de amplitude van de positieve delen van v_i t.g.v. de gelijkrichtdiode.

Als C zich niet te snel kan ontladen (slechte keuze tov. ω_i) volgt v_o de amplitude van v_i niet meer (=diagonaal afknijpen)

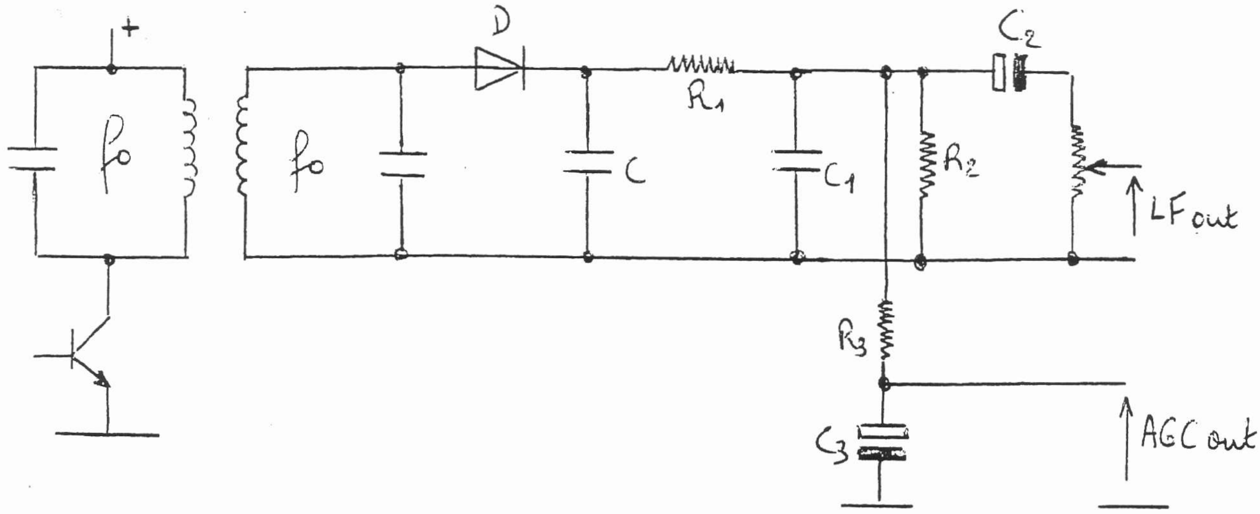


Men bewijst dat een goede keuze gegeven wordt als

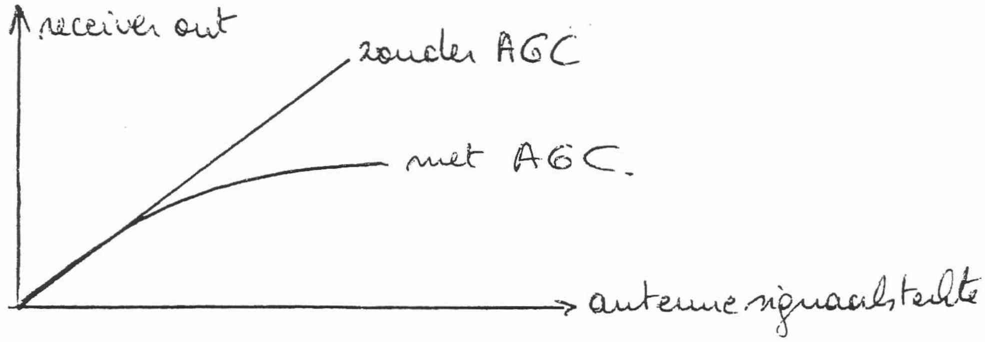
$$RC < \frac{1}{m\,\omega_i}\sqrt{1-m^2}$$

met m de modulatieindeks k.Em.
Anderzijds moet, daar de kondensator zich ook niet te snel mag ontladen: $RC \gg \frac{1}{\omega_C}$

In een ontvanger ziet de detektor in zijn totale gedaante er als volgt uit:



Via de (laatste) middenfrekwentversterker komt het HF signaal terecht op de 2de afgestemde kring op fo (= I.F.). Via diode D wordt het signaal gelijkgericht (positieve alternantie), terwijl C en de rest van de belasting zorgt voor het "gepast volgen" van de LF informatie. Het filter R_1C_1 zorgt voor het verwijderen van de resterende HF komponenten, terwijl R_2 voorzien werd om een DC ontlaadpad voor C te voorzien.
C_2 funktioneert als koppelkondensator zodat over de potentiometer een zuiver LF-signaal overgehouden wordt.
R_3C_3 vormen een LPF zodat audiofrekwenties weggefilterd worden en een DC variërende spanning ontstaat die evenredig is met de sterkte van de draaggolf. Op deze wijze kan men de versterking van de I.F. versterkers aanpassen zodat een nagenoeg konstant signaal verkregen wordt.



4.1.4.1.2 Vermenigvuldiging (synchrone detektie)

Het gemoduleerde signaal is van de gedaante

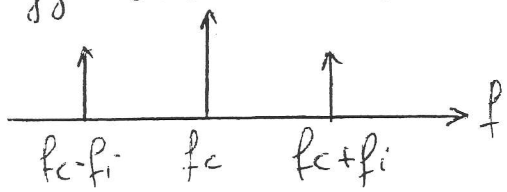
$$\cos\omega_c t + \frac{kE}{2}\cos(\omega_c+\omega_i)t + \frac{kE}{2}\cos(\omega_c-\omega_i)t$$

Vermenigvuldigt men nu dit signaal met $\cos\omega_c t$, dan wordt

$$v_0 = 1/2 + 1/2\cos 2\omega_c t + \frac{kE}{4}\left[\cos\omega_i t + \cos(2\omega_c-\omega_i)t\right]$$

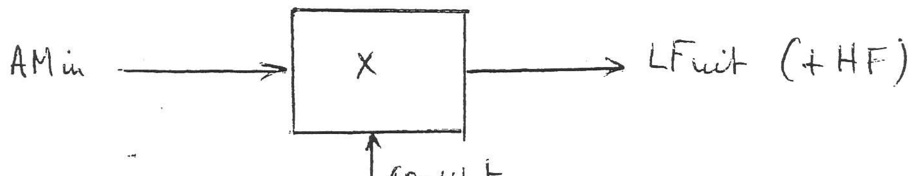
$$+ \frac{kE}{4}\left[\cos\omega_i t + \cos(2\omega_c+\omega_i)t\right].$$

Men verkrijgt dan een verandering van het spektrum van

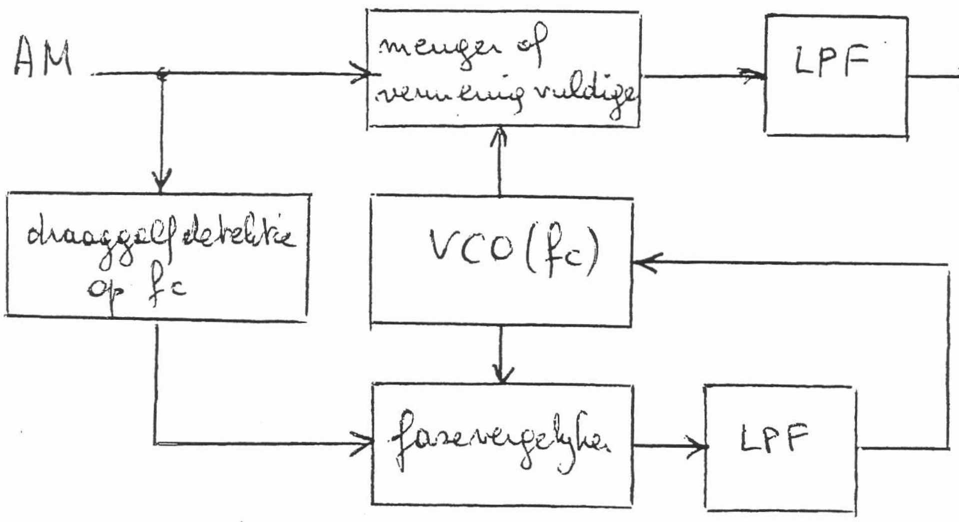


naar

We houden bijgevolg een DC term, een LF term op ω_i en een HF modulatie (AM) op $2\omega_i$ over.
Met een LPF, of vanzelf door de LPF-werking van de volgende audiotrappen, scheidt men de LF-term af.



De vermenigvuldiging kan gebeuren d.m.v. een IC-vermenigvuldiger of m.b.v. een niet lineair element.
Daar er een vaste faserelatie tussen de AM-golf en de trilling $\cos\omega_c t$ dient te bestaan, gebruikt men hiervoor de draaggolf van het AM-signaal zelf, die via een speciale schakeling (Phase locked loop – PLL) afgezonderd kan worden.

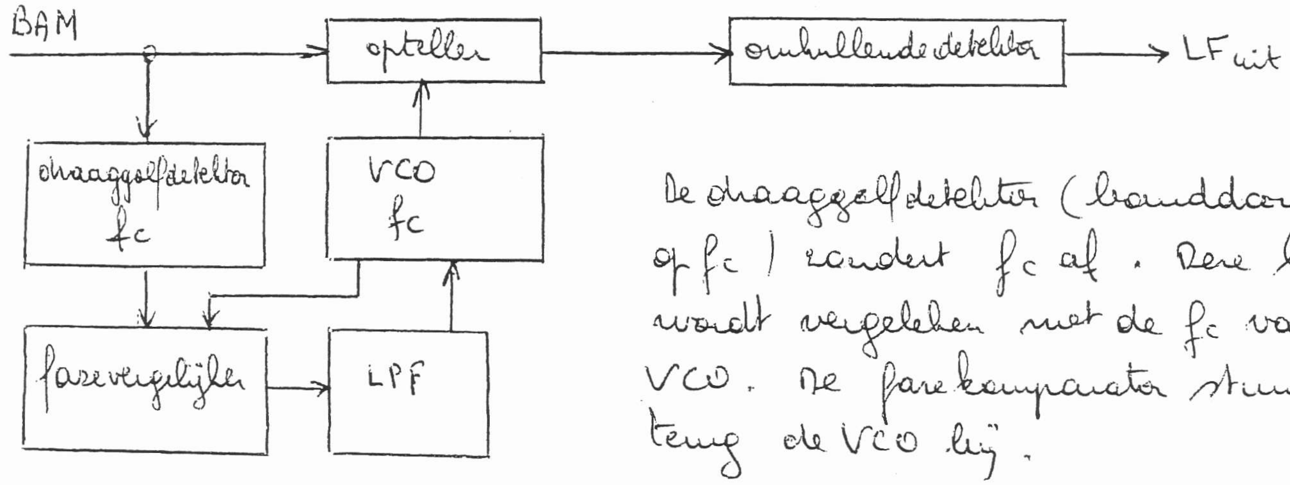


De fasevergelijker vergelijkt ω_c van de eigenlijke draaggolf en van de eigen VCO.
Resultaat van die vergelijking is een DC variërende spanning, maatgevend voor het verschil, die de VCO bijstuurt zodat de vermenigvuldiger steeds de "echte" ω_i aangeboden krijgt.

1.1.4.2. BAM (DSSC)

1.1.4.2.1. Omhullende detektie

Om een omhullende detektor te kunnen toepassen dient men het draaggolfsignaal bij het BAM signaal terug op te tellen, zodat een gewoon AM-signaal verkregen wordt, dat verder via de gewone AM-omhullende detektor kan behandeld worden.
Het is verder van belang dat het fase verband tussen het BAM-signaal en ω_c gerespekteerd wordt (d.w.z. in fase of 180° gedraaid).
Daar men bij BAM in de praktijk steeds een verzwakte drager meezendt, het is immers onmogelijk in de BAM modulator het HF signaal volledig te onderdrukken, zal men in de ontvanger een draaggolf opwekken, afgeleid uit het HF. antennesignaal m.b.v. een PLL.
We verkrijgen volgend blokschema:



De draaggolfdetektor (banddoorlaat op f_c) zondert f_c af. Deze laatste wordt vergeleken met de f_c van de VCO. De fasekomparator stuurt terug de VCO bij.

1.1.4.2.1. vermenigvuldiging (synchrone detektie)

Het BAM signaal wordt vermenigvuldigd met $\cos \omega_c t$, zodat de volgende golfvorm ontstaat:

$$\frac{1}{2} kE \cos \omega_i t + \frac{kE}{4} \cos(2\omega_c - \omega_i)t + \frac{kE}{4} \cos(2\omega_c + \omega_i)t$$

De HF-termen worden terug weggefilterd
Men maakt terug meestal gebruik van een PLL.

1.1.4.3. SSB

Detektie van SSB-signalen is terug mogelijk door synchrone detektie. Het is de enige methode die toepasselijk is.
Voor een LSB signaal verkrijgt men na detektie

$$\frac{1}{2} + \frac{1}{2} \cos 2\omega_c t + \frac{kE}{4} \cos \omega_i t + \frac{kE}{4} \cos(2\omega_c - \omega_i t)$$

Met een LPF kan men terug de nuttige komponent op ω_i afzonderen.

1.1.4.4. SSSC

De synchrone detektie is terug de enige detektiemogelijkheid
Na demodulatie heeft men

$$\frac{kE}{4} \cos \omega_i t + \frac{kE}{4} \cos(2\omega_c - \omega_i)t$$

Men laat de demodulator terug uitgeven op een LPF.

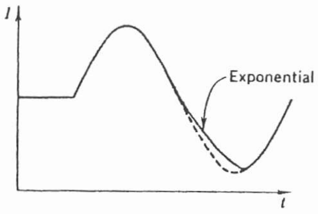


Fig. 7-15 Diagonal clipping.

to follow the change. As a result, the currrent will decay exponentially, as shown in Fig. 7-15, instead of following the waveform; this is called *diagonal clipping*. It does not normally occur when percentage modulation (at the highest modulation frequency) is below about 60 percent, so that it is possible to design a diode detector that is free from this type of distortion. Nevertheless, one should still be aware of its existence as a limiting factor on the size of the RF filter capacitors.

1.1.5. 7-3 COMMUNICATIONS RECEIVERS

A communications receiver is one whose main function is the reception of signals used for communications rather than for entertainment. It is a radio receiver designed to perform the tasks of low- and high-frequency reception better than the type of set found in the average household. In turn, this makes the communications receiver useful in other applications, such as the detection of signals from high-frequency impedance bridges (where it is used virtually as a high-sensitivity tuned voltmeter), signal-strength measurement, fairly accurate frequency measurement, and even detection and display of individual components of a high-frequency wave (such as an FM wave with its many sidebands). It is often operated by electronically qualified people, so that any added complications in its tuning and operation are not necessarily detrimental, as they would have been in a receiver to be used by the general public.

The communications receiver is similar in many respects to the ordinary home receiver, as the block diagram of Fig. 7-16 and the photograph of Fig. 7-17 demonstrate. Both are, for example, superheterodyne receivers, but in order to perform its tasks the communications receiver has a number of modifications and added features. These are the subject of this section, in which the strange new blocks of Fig. 7-16 will also be treated.

7-3.1 Extensions of the Superheterodyne Principle

Whereas some of the circuits found in communications receivers, such as tuning indicators and beat-frequency oscillators, may be said to be mere additions, other circuits appear to extend the superheterodyne

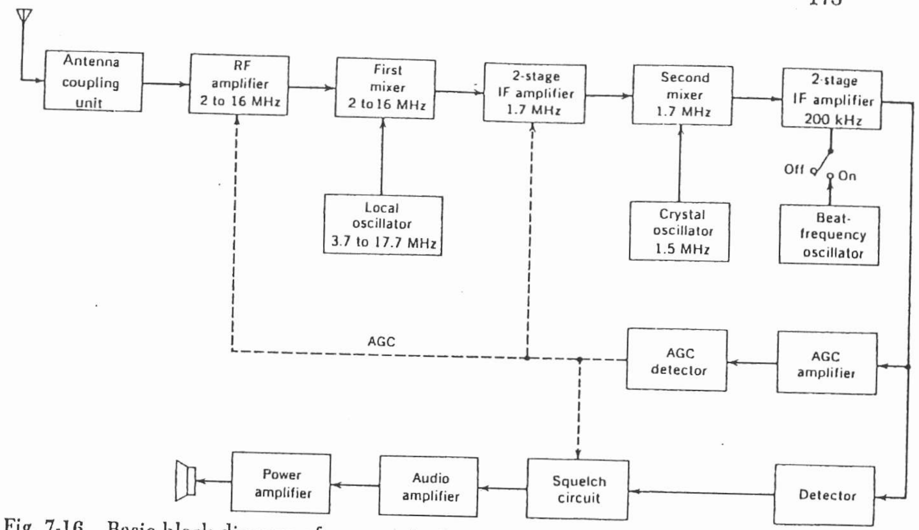


Fig. 7-16 Basic block diagram of communications receiver.

principle further. Delayed AGC and double conversion are but two of these circuits. It has thus been found convenient to subdivide the topic into extensions of the superheterodyne principle on the one hand, and additions to it on the other.

Input stages It is common to have one, or sometimes even two, stages of RF amplification. Two stages are preferable if extremely high sensitivity and low noise are required, although some complications in tracking are bound to occur. Regardless of the number of input stages, some system of band changing will have to be used if the receiver is to cover

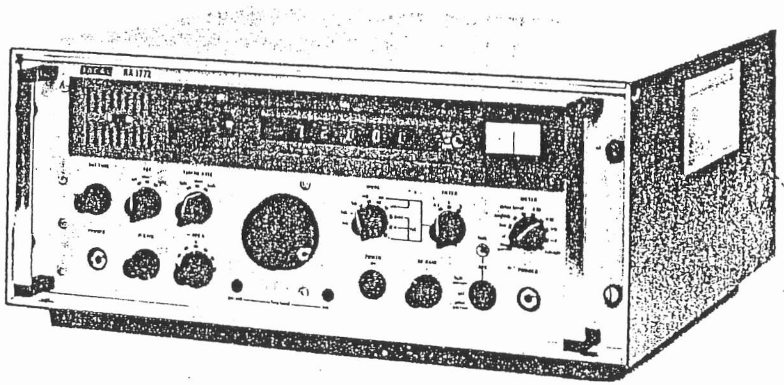


Fig. 7-17 Communications receiver. (*Courtesy of Racal Electronics Pty. Ltd.*)

a wide frequency range, as nearly all communications receivers do. This is compounded by the fact that the normal variable capacitor cannot be relied upon to cover a frequency ratio much in excess of 2:1 at high frequencies. Band changing is accomplished in either of two ways: by switching in the required RF, mixer, and local oscillator coils, or by frequency synthesis.

In order to obtain maximum efficiency from different antenna systems, or at different frequencies, provision is made in many good-quality communications receivers for matching various antenna input impedances. For this purpose, different sockets and trimmers, tapped transformers or even whole matching networks may be provided (see also Sec. 10-5). The coupling network, if adjustable, is not normally meant to be continuously tunable, but it is simply tuned for optimum results in the middle of each band.

Bandspread A bandspread control is an essential adjunct of communications receivers. As the name implies, bandspreading permits stations transmitting on frequencies very close to each other to be resolved by the receiver. This is achieved by increasing the physical distance between them on the dial, or by providing a subsidiary dial on which they can be separated. Either mechanical or electrical means may be used to provide bandspread.

In the mechanical system, the bandspread control is geared to the main tuning control. The gearing is made such that the fine control is very similar to a vernier, and one turn of the main control corresponds to several turns of the fine tuning. In one such commercial receiver, the fine-tuning mechanism is gear-driven, and the bandspread reduction is 140:1. The receiver of Fig. 7-17 produces the same results with synthesis and a digital frequency display as shown. Some provision must be made for the disconnection of the mechanical type of bandspread to permit rapid access from one end of the dial to the other.

In the electrical bandspread system, the ganged capacitor is shunted by a ganged trimmer, which may give a variation of 30 pF for a full revolution where the main tuning control gives 300 pF. The close stations are separated once again, but this time on a separate dial. Mechanical bandspread is quite common in current receivers, electrical bandspread is in decline, and frequency synthesis is very much on the upswing.

Double conversion Comunications receivers, and some high-quality domestic AM receivers, have more than one intermediate frequency—generally two, but sometimes even more. When a receiver has two different IFs, as does the one shown in block form in Fig. 7-16, it is then said to be a *double-conversion* receiver. The first IF is high, generally several megahertz, and the second one quite low, of the order of 200 kHz

2 IF stages

or even less. After leaving the RF amplifier, the signal in such a receiver is still mixed with the output of a local oscillator. This is similar in all respects to the local oscillator of a domestic receiver, except that now the resulting frequency difference is a good deal higher than the usual 455 kHz. The high intermediate frequency is then amplified by the high-frequency IF amplifier, and the output is fed to a second mixer and mixed with that of a second local oscillator. Since the second local oscillator frequency is normally fixed, this could be a crystal oscillator, and in fact very often is, in nonsynthesized receivers. The low second intermediate frequency is amplified by an LF IF amplifier, and then detected in the usual manner.

Double conversion is essential in communications receivers. As will be recalled from Sec. 7-2.3, the intermediate frequency selected for any receiver is bound to be a compromise since there are equally compelling reasons why it should be both higher and lower. Double conversion avoids this compromise. The high first intermediate frequency pushes the image frequency further away from the signal frequency, and therefore permits much better attenuation of it. The low second IF, on the other hand, has all the virtues of a low fixed operating frequency, particularly sharp selectivity and hence good adjacent-channel rejection.

Please note that the *high intermediate frequency must come first*. If this does not happen, the image frequency will be insufficiently rejected at the input and will become inextricably mixed with the proper signal, so that no amount of high IF stages will make any difference afterward.

The result of having two such intermediate frequencies is that double-conversion receivers provide a combination of higher image and adjacent-frequency rejection than can be achieved with the simple superheterodyne system. It should be noted, on the other hand, that double conversion offers no great advantages for broadcast or other medium-frequency receivers. However, it is essential for receivers operating in the crowded short-wave bands. See also Sec. 7-5.1 and Ref. 2 for a further discussion of double conversion and its achievement with novel techniques.

Delayed AGC Simple AGC, as treated in Sec. 7-2.4, is clearly an improvement on no AGC at all, in that the gain of the receiver is reduced for strong signals. Unfortunately, as Figs. 7-13 and 7-18 both show, even weak signals do not escape this reduction. Figure 7-18 also shows two other AGC characteristics. The first is the "ideal" characteristic. In this no AGC is applied until signal strength is considered adequate, and after this point a constant average output is obtained no matter how much more the signal strength rises. The second is the *delayed* AGC curve. This shows that AGC bias is not applied until the signal strength has reached a predetermined level, after which bias is applied as with normal AGC, but more strongly. As the signal strength then rises, re-

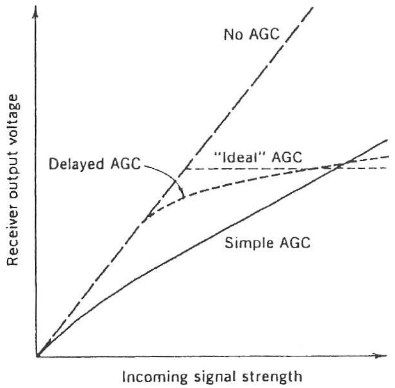


Fig. 7-18 Various AGC characteristics.

ceiver output also rises, but relatively slightly. The problem of reducing the gain of the receiver for weak signals has thus been avoided, as with "ideal" AGC.

A very common method of obtaining delayed AGC is shown in Fig. 7-19. It uses two separate diodes: the detector and the AGC detector. These can be connected either to separate transformer windings as shown, or both to the secondary without too much interference. As indicated, a positive bias is applied to the cathode of the AGC diode, to prevent conduction until a predetermined signal level has been reached. A control is often provided, as shown, to allow manual adjustment of the bias on the AGC diode, and hence of the signal level at which AGC is applied. If mostly weak stations are likely to be received, the delay control setting may be quite high (i.e., no AGC until signal level is fairly high). Nevertheless, it should be made as low as possible, to prevent overloading of the last IF amplifier by unexpected stronger signals.

The method just described works well with FETs, and also with bipolar transistors if the number of stages controlled is large enough. If in the latter case fewer than three stages are being controlled, it may not be possible to reduce the gain of the receiver sufficiently for very strong signals, because of collector leakage current. If that is so, a secondary method of AGC is sometimes used together with simple AGC, the overall result being not unlike delayed AGC. A diode is here employed for variable damping, in a manner akin to that used in the ratio detector, as described in Sec. 7-4.4.

Variable sensitivity and selectivity The ratio of the highest to the lowest signal strengths which a communications receiver may have to cope with could be as high as 10^5:1. This means that the receiver must have sufficient sensitivity to amplify fully very weak signals, and it must also be capable of having its gain reduced by AGC action by a ratio of 10^5:1, or 100 dB, so as not to overload on the strongest signal. Even the best

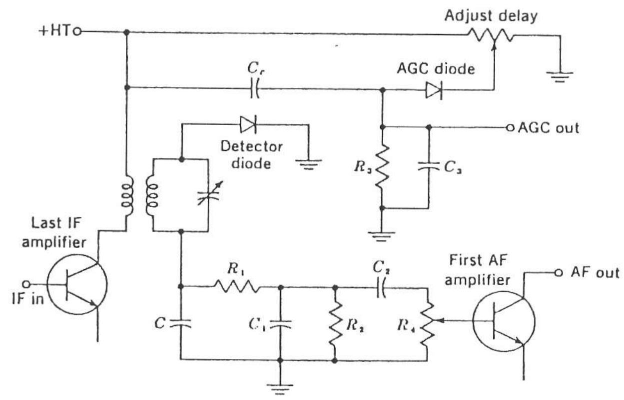


Fig. 7-19 Delayed AGC circuit.

AGC system is not capable of this performance. Apart from the alarming variations in output that may occur, there is also the risk of overloading several of the IF amplifiers, especially the last one, and also the demodulator diode. To prevent the distortion which would follow, and also possibly the permanent damage, the most sensitive communications receivers incorporate a sensitivity control.[1] This generally consists of a potentiometer which varies the bias on the RF amplifier, and is, in fact, an RF gain control. The AGC is still present, but it now acts to keep the sensitivity of the receiver to the level determined by the setting of the potentiometer. The receiver is now considerably more versatile in handling varying input signal levels.

The selectivity, or, to be more precise, the bandwidth, of the low-frequency IF amplifier may be made variable over a range that is commonly 1 to 12 kHz. The largest bandwidth permits reception of high-quality broadcasts, whereas the smallest (although it greatly impairs this quality) reduces noise and therefore increases intelligibly, and will also reduce adjacent-channel interference. Variable selectivity is achieved in practice by switching in (noninductive) resistors across the primary and secondary of the last LF IF transformer. For instance, if this IF is 110 kHz, $Q_L = (\sqrt{2} \times 110)/1 = 155$ for a bandwidth of 1 kHz. This value is quite feasible, of course. A set of resistors is provided, any of which may be switched across the tank to give bandwidths of (say) 2, 4, 6, 8, 10, and 12 kHz. Alternatively, a crystal filter may be used in a similar manner to provide the narrower bandwidths. Receivers designed for radiotelegraphy reception may have minimum bandwidths as low as 300 Hz.

A *notch filter* is sometimes found in a communications receiver. This

[1] Unfortunately, this statement does not work in reverse. Mere possession of a sensitivity control by a receiver does not guarantee that it is, in fact, a *sensitive* receiver.

is a wavetrap, or a stop filter, designed to reduce receiver gain at some specific frequency and therefore help to reject it. It often consists simply of a series-resonant circuit across one of the LF IF transformers. The frequency at which this trap is resonant will naturally be rejected since the load impedance of that amplifier will then be almost short-circuited. If the capacitor in the series-resonant circuit is made variable, the position of the notch can be adjusted so that any one adjacent spurious signal may be rejected on either side of the IF passband. A crystal gate (cf. Sec. 3-1.2) may be used similarly. The versatility of the receiver has naturally been enhanced, since it now has the notch filter, variable selectivity and double conversion for suppressing unwanted nearby signals.

Blocking If a radio receiver is tuned to a weak signal, naturally the developed AGC will be low and the front-end gain high. If a strong signal not too distant in frequency is now received, then unless it is properly rejected, it could develop substantial AGC voltage. Such a high AGC, caused by a spurious signal, could reduce the gain of the receiver, perhaps to the point of making the wanted signal inaudible. This situation is unwelcome and, if the interfering signal is intermittent, it is intolerable. A receiver whose AGC system has very little reaction to the nearby spurious signals is said to have good *blocking*. A good way of showing how blocking is defined and measured is to state how it is quoted in receiver specifications. The Redifon R 551 is a receiver with very good blocking performance, quoted by the manufacturers as follows: "With a 1 mV EMF A0 (SSB, 1000 Hz tone) wanted signal, a simultaneous 6 V EMF A0 unwanted signal (at least 20 kHz from wanted signal) will not reduce the wanted AF output by more than 3 dB."

Needless to say, very high IF rejection of adjacent signals is needed to produce such excellent blocking performance. Yet this performance is required in SSB receivers, and all other instances of working in crowded frequency bands.

7-3.2 Additional Circuits

Whereas the foregoing circuits and characteristics were most easily classified as extensions of the superheterodyne system, the following are best thought of as additions. It must however be admitted that the subdivision, although convenient, is at times a little artificial.

Tuning calibration This consists of having a built-in crystal oscillator, usually operating at 500 to 1000 kHz, whose output may be fed to the input of the receiver by throwing the appropriate switch. With the *beat-frequency oscillator* in operation (to follow), whistles will now be heard

at 500- or 1000-kHz intervals, especially since the crystal oscillator works into a resistive load, so as not to attenuate harmonics of the fundamental frequency. The calibration of the receiver may now be corrected by adjustment of the pointer or cursor, which must, of course, be movable independently of the gang. An elaborate receiver, which is also tunable to freqencies above 30 MHz, may have a built-in crystal amplifier, whose function is to amplify the higher harmonics of the crystal oscillator to make frequency calibration easier at those frequencies. Synthesized receivers do not require this facility.

Beat-frequency oscillator (BFO) A communications receiver should be capable of receiving transmissions of Morse Code, i.e., pulse-modulated RF carrier. In the diode detector of a normal receiver, since there is no provision for registering the difference between the presence and the absence of a carrier,[1] such pulse-modulated dots, dashes and spaces would produce no output whatever from the detector.

In order to make Morse Code audible, the receiver has a built-in beat-frequency oscillator, normally at the detector, as shown on the block diagram of Fig. 7-16. The BFO is not really a beat-frequency oscillator at all; it is merely a simple *LC* oscillator. The Hartley BFO is one of the favorites, operating at a freqency of 1 kHz or 400 Hz above or below the last intermediate frequency. When the latter is present, a whistle is heard in the loudspeaker, so that it is the combination of the receiver, detector, input signal and this extra oscillator which has now become a beat-frequency oscillator. Since signal is present only during a dot or a dash in Morse Code, only these are heard; thus the code can be received satisfactorily, as can radiotelegraphy. To prevent interference, the BFO is switched off when normal reception is resumed.

Noise limiter A fair proportion of communications receivers are provided with noise limiters. The name is a little misleading since it is patently not possible to do anything about random noise in an AM receiving system (it is possible to reduce random noise in FM, as will be seen). Such a noise limiter is really an *impulse-noise limiter*, a circuit for eliminating, or at least reducing, the interfering noise pulses created by ignition systems, electrical storms or electrical machinery of various types. This is often done by automatic silencing of the receiver for the duration of a noise pulse, which is preferable to a loud, sharp noise in the loudspeaker or headphones. In a common type of noise limiter, a diode is used in conjunction with a differentiating circuit. The limiter circuit provides a negative voltage as a result of the noise impulse or any very

[1] This is not strictly true since there are two ways of doing this, but neither is satisfactory for Morse Code and dial calibration. First, there is the fact that noise comes up strongly when the carrier disappears, and second, a signal-strength meter or tuning indicator would show the presence of a carrier, but much too slowly.

sharp voltage rise, and this negative voltage is applied to the detector, which is thus cut off. The detector then remains cut off for the duration of the noise pulse, a period that generally does not exceed a few hundred milliseconds. It is essential to provide a facility for switching off the noise limiter, or else it will interfere with Morse Code or radiotelegraphy reception.

There are many different types of noise limiters, all used to suppress impulse noise; cf. Ref. 3.

Squelch (muting) When no carrier is present at the input of a sensitive receiver, i.e., in the absence of transmissions on a given channel or between stations, a sensitive receiver will produce a disagreeable amount of noise. This is because AGC disappears in the absence of any carrier, the receiver acquires its maximum sensitivity, and amplifies the noise present at its input. In many circumstances this is not particularly important, but in many others it can be annoying and tiring. Systems such as those used by the police, ambulances, and coast radio stations, in which a receiver must be tuned and manned at all times but transmission is sporadic, are the principal beneficiaries of squelch. This enables the receiver's output to remain cut off unless the carrier is present. Apart from eliminating inconvenience, such a system must naturally increase the efficiency of the operator. Squelch is also called *muting* or *quieting*. Quiescent (or quiet) AGC and Codan (carriers operated device, anti-noise) are similar systems.

The squelch circuit, as shown in Fig. 7-20, consists of a dc amplifier to which AGC is applied and which operates upon the first audio amplifier of the receiver. When the AGC voltage is low or zero, the dc amplifier, T_2, draws current so that the voltage drop across its load resistor R_1 cuts off the audio amplifier, T_1; thus no signal or noise is passed. When the AGC voltage becomes sufficiently negative to cut off T_2, this dc amplifier no longer draws collector current, so that the only bias now on T_1 is its self-bias, furnished by the bypassed emitter resistor R_2 and also by the base potentiometer resistors. The audio amplifier now functions as though the squelch circuit were not there.

R_3 is a dropping resistor, whose function it is to ensure that the high tension supplied to the collector and base potentiometer of T_1 is higher than the high tension supplied (indirectly) to its emitter. Manual adjustment of R_3 will allow the cut-in bias of T_2 to be varied so that quieting may be applied for a range of selected values of AGC. This facility must be provided, otherwise weak stations, not generating sufficient AGC, might be cut off. The squelch circuit is normally inserted immediately after the detector, as shown in Figs. 7-16 and 7-20.

Automatic frequency control As will be recalled from Sec. 5-3.3, the heart of an AFC circuit is a frequency-sensitive device, such as the phase

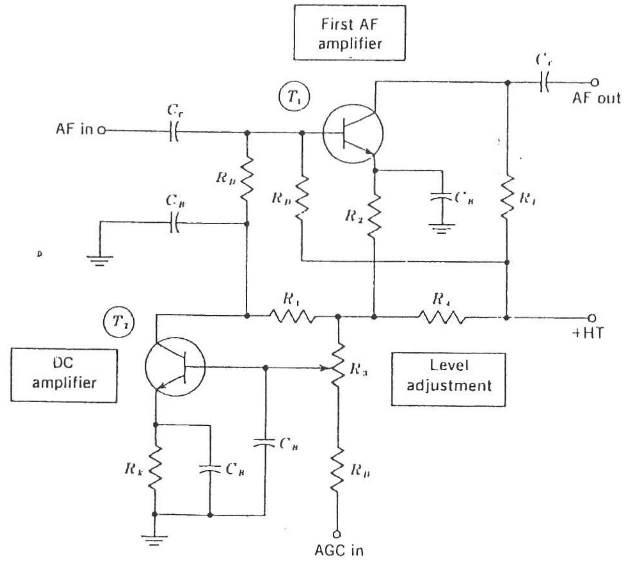


Fig. 7-20 Typical squelch circuit.

discriminator, which produces a dc voltage whose amplitude and polarity are proportional to the amount and direction of the local oscillator frequency error. This dc control voltage is then used to vary, automatically, the bias on a variable-reactance device, whose output capacitance is thus changed. This variable capacitance appears across the (first) local oscillator coil, and (in the manner described in Sec. 5-3.3) the frequency of this VFO* is automatically kept from drifting with temperature, line voltage changes or component aging. A block diagram of a receiver AFC system is shown in Fig. 7-21.

It is worth noting that the number of extra stages required to provide AFC is much smaller in a double-conversion receiver than in the stabilized reactance modulator, since most of the functions required are already present. On the other hand, not all receivers require AFC, especially not synthesized ones. Those that benefit most from its inclusion are undoubtedly SSB receivers, whose local oscillator stability must be exceptionally good to prevent drastic frequency variations in the demodulated signal.

Metering A built-in meter with a function switch is very often provided. It is very helpful in diagnosing any faults that may occur, since it measures voltages at key points in the receiver. One of the functions (sometimes the sole function) of this meter is to measure the incoming signal

* Variable-frequency oscillator, a commonly used term in such a situation.

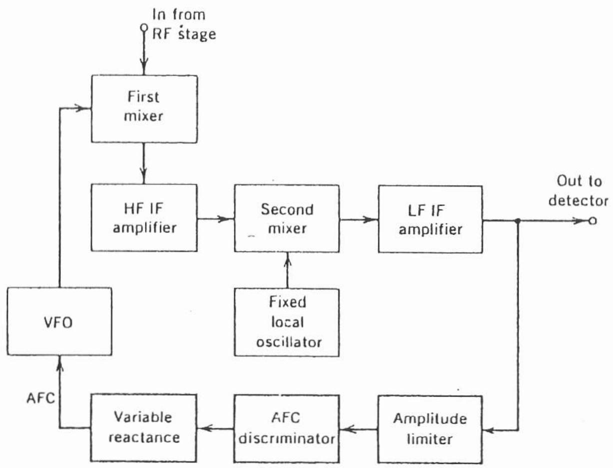


Fig. 7-21 Block diagram of receiver with AFC.

strength. It is then called an S *meter,* and very often reads the collector current of an IF amplifier to which AGC is applied, as shown in Fig. 7-22. Since this collector current decreases as the AGC goes up, the meter has its zero on the right-hand side. The S meter may sometimes be in an unbalanced bridge and hence forward-reading. In either case, the calibration of the meter is likely to be quite arbitrary because of the great variation of the sensitivity of the receiver through the bands, especially if there is a sensitivity control or adjustable delayed AGC.

A receiver with an S meter is more versatile than one without, not only because tuning to a wanted signal can now be more accurate, but also because the receiver can now be used as a relative signal-strength meter and also as the detector for an RF impedance bridge. It can also be used for applications such as tuning individually to the various sideband frequencies of an FM signal. This can determine the presence of those components and demonstrate the disappearance of the carrier for certain values of modulation index, from which readings deviation and linearity of the FM source may be determined (see Sec. 5-1.3).

FM and SSB reception Some receivers have provision for the reception of FM, either the narrowband FM used by mobile networks or the high-quality broadcast transmissions in the 88- to 108-MHz band. To allow FM reproduction, a receiver requires broadband IF stages, an FM demodulator and an amplitude limiter; these are described later in this chapter.

More and more present-day communications receivers have facilities for single-sideband reception. Basically this means that a product detector (see Sec. 7-5) must be provided, but it is also very helpful if there

is an AFC system present, as well as variable selectivity (preferably with a crystal filter), since the bandwidth used for SSB is narrower than for ordinary AM.

Diversity reception This is not so much an additional circuit in a communications receiver as a specialized method of using such receivers. There are two forms: *space diversity* and *frequency diversity.*

Whereas AGC helps greatly to minimize some of the effects of fading, it cannot help when the signal fades into the noise level. Diversity-reception systems make use of the fact that although fading may be severe at some instant of time, some frequency, and some point on earth, it is extremely unlikely that signals at different points or different frequencies will fade simultaneously. (See also Sec. 9-2.2 for a detailed description of fading, its various causes and its effects upon reception.)

Both systems are in constant use, by communications authorities, commercial point-to-point links and the military. In space diversity, two or more receiving antennas are employed, separated by nine or more wavelengths. There are as many receivers as antennas, and arrangements are made to ensure that the AGC from the receiver with the strongest signal at the moment cuts off the other receivers. Thus only the signal from the strongest receiver is passed to the common output stages.

Frequency diversity works in much the same way, but now the same antenna is used for the receivers, which work with simultaneous transmissions at two or more frequencies. Since frequency diversity is more wasteful of the frequency spectrum, it is used only where space diversity cannot be employed, such as in restricted spaces where receiving antennas could not have been separated sufficiently. Ship-to-shore and ship-to-ship communications are the greatest users of frequency diversity at HF.

As described, both systems are known as *double-diversity* systems, in that there are two receivers employed in a diversity pattern. Where conditions are known to be critical, as in *tropospheric scatter* communications, *quadruple diversity* is used. This is a space-diversity system which has receiver arrangements as just described, with two transmitters

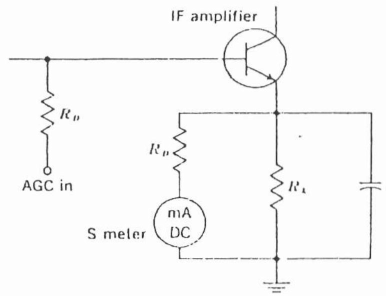


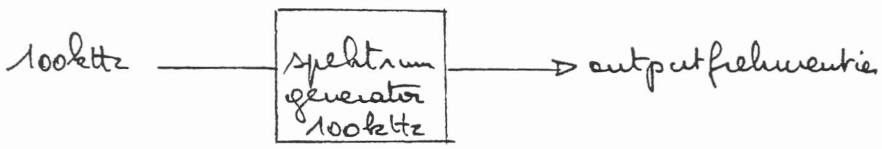
Fig. 7-22 S meter.

Opmerking: frekwentie synthesizers

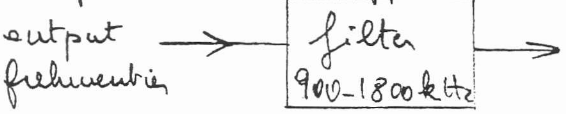
Ontvangers moeten een variabele lokale oscillator hebben om de verschillende frekwenties te kunnen ontvangen.
Om stabiele frekwenties te verkrijgen zou dit een kristal per frekwentie vereisen.
De eerste frekwentie synthesizers werkten dan ook volgens dit principe: de multipele - kristal synthesizer genoemd. Er waren zoveel oscillators als frekwentiedekades, en iedere oscillator was voorzien van 10 kristallen. De vereiste frekwentie werd geselekteerd door het gepaste kristal in iedere oscillator te schakelen, waarna de synthesizer de uitgangen mengde om de gewenste frekwentie te selekteren.
Een nadeel van deze metode is dat een stabiele frekwentie verscheidene kristalovens vereist
Tegenwoordig werkt men met slechts 1 (zeer stabiele) kristal oven die 1 kristal bevat.

Direkte synthesizer

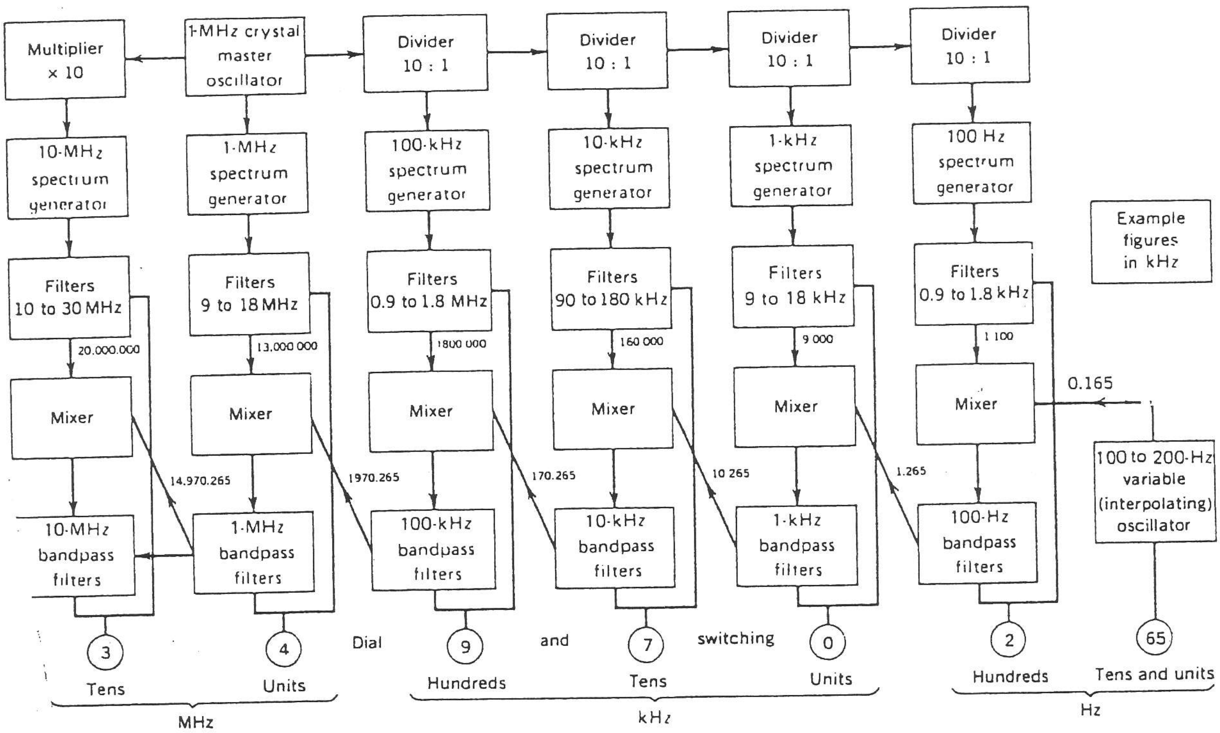
Hier wordt er gewerkt met 1 uiterst stabiele kristaloscillator die in een kristaloven geplaatst wordt. (master oscillator)
Uit deze oscillator worden een aantal nevenfrekwenties afgeleid door frekwentie delen en vermenigvuldigen ($\times 10, \div 10^1, \cdots \div 10^4$)
Deze verschillende frekwenties gaan uit op spektrumgenerators die een veelvoud (harmonische) kunnen geven van deze nevenfrekwenties.
Zo zal een inputfrekwentie van 100 kHz outputfrekwenties van 200 kHz, 300 kHz, ..., 900 kHz, ---, 1800 kHz, 1900 kHz ... afgeven.



De outputfrekwenties gaan uit op een filter van 900 kHz - 1800 kHz, afstembaar in stappen van 100 kHz.



Door nu het ingestelde uitgangssignaal te mengen met frekwenties van de vorige dekades (mengen is hier frekwenties optellen) kan men in principe om 't even welke frekwentie selekteren.
Het totale blokschema is hieronder gegeven.



Om bijvoorbeeld de ingestelde frekwentie van 34 970,265 kHz te selekteren gaat men als volgt tewerk:

<u>eenheden en tientallen</u>: De te realiseren frekwentie is 65 Hz. Men stelt nu een (uiterst stabiele RC)-oscillator in op 165 Hz, dus 10 honderdtallen meer dan de 'echte' frekwentie, of

$$f_{RC} = f_{<100Hz} + 100.$$

Het signaal f_{RC} wordt dan aangeboden aan de volgende trap.

<u>honderdtallen t.e.m. voorlaatste dekade</u>

Hier stelt men iedere oscillator in op de te realiseren digit + 9 eenheden, dwz: 2 wordt 200 + 900 = 1100 Hz

0 wordt 0+9000 = 9000 Hz

7 wordt 70.000 + 90.000 = 160 kHz

9 wordt 900 kHz + 900 kHz = 1800 kHz

4 wordt 4 MHz + 9 MHz = 13 MHz

Deze frekwenties worden telkens gemengd (= opgeteld) met de frekwentie van de vorige dekade, en vooraf gefilterd met een stappenfilter (100 Hz - 1000 Hz - ····· - 1 MHz -stap), die op het voorpaneel d.m.v. schakelaars ingesteld worden. Dit zijn dus banddoorlaatfilters.

Voor de honderdtallen heeft men dan 1100 + 165 = 1265 Hz

⋮

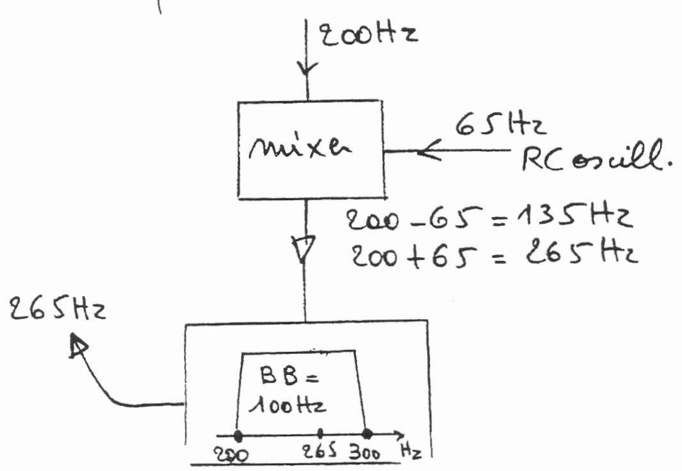
Voor de M-Hertzen heeft men dan 13 M + 1,970265 M = 14,970265 MHz

<u>laatste dekade</u>: Moest men nu de 10 MHz-dekade instellen op 30 M, dan zou 10 MHz teveel afgegeven worden.

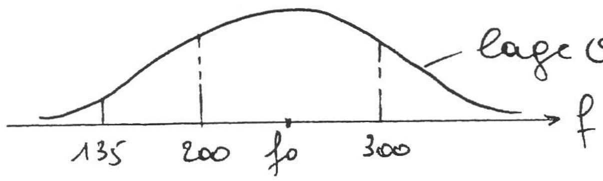
Daarom stelt men de hoogste dekade in op het digit − 1 dus 3 − 1 => 20 MHz.

Deze 20 MHz wordt dan gemengd met de vorige dekade (14,970265 MHz), waardoor de gevraagde 34.970.265 Hz ontstaat.

De reden dat men er zo'n ingewikkeld frekwentieplan op na houdt is de volgende: Beschouw de honderdtallen, de <u>band-breedte</u> van het filter is 100 Hz, onafhankelijk van de te realiseren frekwentie. Moesten we rechtstreeks 265 Hz wensen te realiseren, dan zou de RC-oscillator 65 Hz en de spektrum generator-filter kombinatie 200 Hz moeten afgeven

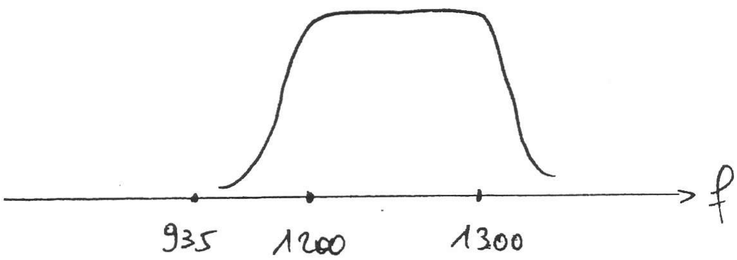


Daar de mixer zowel som als verschilfrekwentie afgeeft (265 en 135 Hz) en de banddoorlaatfilter signalen van 200 tot 300 Hz moet doorlaten om de 265 Hz af te geven aan de volgende dekade, en terzelfdertijd de 135 Hz de pas moet afsnijden, zou dit een te brede doorlaatband vergen t.o.v. de centerfrekwentie $\sqrt{300 \cdot 200}$ Hz, waardoor de Q-faktor laag zou moeten genomen worden met als gevolg het toch doorlaten van de 135 Hz



$$Q = \frac{2\pi f_0 L}{R}$$

Door echter op de hierboven beschreven manier te werken heeft men voor de te realiseren frekwentie een BPF nodig met zelfde bandbreedte (100 Hz), doch op een veel hogere frekwentie (1200-1300 Hz). De verschilfrekwentie 1100-165 = 935 Hz kan nu wel voldoende verzwakt worden door gebruik te maken van een hoge Q-faktor (of een Butterworth of Chebyshev filter met lage rimpel)

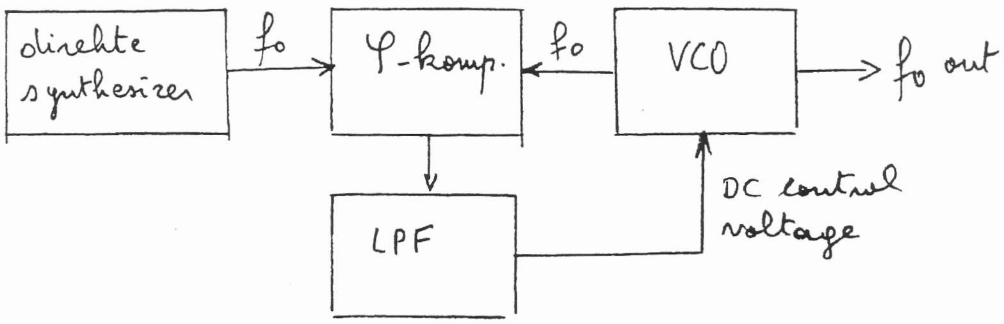


Vandaar het komplekse frekwentieplan.

Als nadeel van deze direkte synthesizer vermelden we het toch doorkomen van verschilfrekwenties en de aanwezigheid van ruis.

indirekte synthesizers:

De indirekte synthesizer is opgebouwd rond een PLL, en wel op de volgende wijze:



De direkte synthesizer wordt als input gebruikt voor de fasekomparator en wordt dus als een soort master-oscillator gebruikt. De fasekomparator vergelijkt deze frekwentie met de frekwentie van de VCO, zodat deze laatste, bij fase afwijkingen, via een DC-kontrole spanning bijgestuurd wordt.

Waarom nu op deze manier tewerk gaan?

- Zoals vermeld geeft de direkte synthesizer veel ruis af, en worden soms toch nog verschilfrekwenties van de mixer doorgegeven. Deze ruissignalen zullen nu, dank zij het volgen van het sterke master-oscillator signaal door de VCO, niet meer doorgegeven worden op fo out. De VCO zal dus enkel de frekwentie fo afgeven (en volgen).
- Het blijkt dat short-term stability bij een kristaloscillator (dus tijdspannes ms) minder goed is dan deze van een goed gedimensioneerde vrijlopende (zonder Xtal) - oscillator. Dit gekoppeld aan de uitermate goede long-term stability van de kristaloscillator, die uiteraard door de VCO gevolgd wordt, resulteert in een uitstekende long- en short-term-stabiliteit.

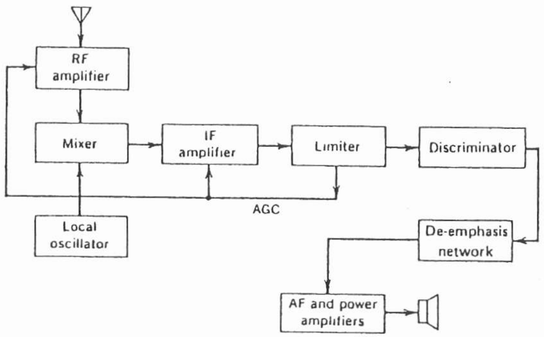


Fig. 7-23 FM receiver block diagram.

at each end of the link arranged just like the receivers. This ensures that signals of adequate quality will be received under even the worst possible conditions. (See Sec. 9-2.4, where tropospheric scatter is described fully and the use of diversity with it is discussed, and also Ref. 7.)

There is one snag, unfortunately, that applies to diversity systems and limits their use in voice communications. Since, in general, each signal travels over a slightly different path, the audio output will have a phase difference when compared with that of the other receiver(s). As a result, diversity reception is used very often for telegraph or data transmission (i.e., pulses), but present diversity systems for voice communications leave much to be desired, unless some form of pulse modulation is employed for the voice transmission (the most popular form is *pulse-code modulation*, as described in Sec. 15-2.4).

1.2. Outvangertypes voor FM.

1.2.1. 7-4 FM RECEIVERS

The FM receiver is a superheterodyne receiver, and the block diagram of Fig. 7-23 shows just how similar it is to an AM receiver. The basic difference are as follows:

1. Much higher operating frequencies in FM.
2. Need for limiting and de-emphasis in FM.
3. Totally different methods of demodulation.
4. Different methods of obtaining AGC.

7-4.1 Common Circuits—Comparison with AM Receivers

A number of sections of the FM receiver correspond exactly to those of other receivers already treated; for example, the same criteria apply in the selection of the intermediate frequency, and IF amplifiers are basically similar. Again, a number of concepts have very similar meanings so that only the differences and special applications need be pointed out.

RF amplifiers An RF amplifier is always used in an FM receiver. The main reason is to reduce the noise figure, which could otherwise be a problem because of the large bandwidths needed for FM. It is also required to match the input impedance of the receiver to that of the antenna. To meet the second requirement, grounded gate (or base) or cascode amplifiers are employed. Both types have the property of low input impedance, matching the antenna, and neither requires neutralization. This is because the input electrode is grounded in either type of amplifier, effectively isolating input from output. A typical FET grounded-gate RF amplifier is shown in Fig. 7-24. It has all the good points mentioned, and the added features of low distortion and simple operation.

Frequency changers The oscillator circuit takes any of the usual forms, with the Colpitts and Clapp predominant, being suited to VHF operation. Tracking is not normally much of a problem in FM broadcast receivers. This is because the tuning frequency range is only 1.25:1, much less than in AM broadcasting.

A very satisfactory arrangement for the front end of an FM receiver consists of FETs for the RF amplifier and mixer, and a bipolar transistor oscillator. As implied by this statement, separately excited oscillators are normally used, with an arrangement as shown in Fig. 7-6.

Intermediate frequency and IF amplifiers Again, the types and operation do not differ much from their AM counterparts. It is worth noting, however, that the intermediate frequency and the bandwidth required are far higher than in AM broadcast receivers. Typical figures for receivers operating in the 88- to 108-MHz band are an IF of 10.7 MHz and a bandwidth of 200 kHz. As a consequence of the large bandwidth, gain per stage may be low. Hence two IF amplifier stages are often provided, in which case the shrinkage of bandwidth as stages are cascaded must be taken into account.

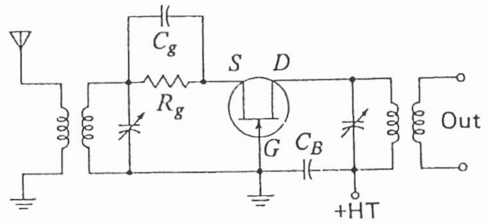


Fig. 7-24 Grounded-gate FET RF amplifier.

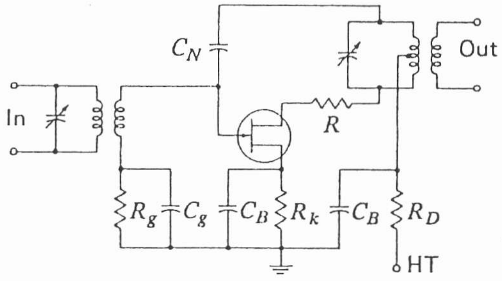


Fig. 7-25 Amplitude limiter.

7-4.2 Amplitude Limiting

In order to make full use of the advantages offered by FM, a demodulator[1] must be preceded by an amplitude limiter, as discussed in Chap. 5. This is on the grounds that any amplitude changes in the signal fed to the FM demodulator are spurious. They must therefore be removed if distortion is to be avoided. The point is significant, since most FM demodulators react to amplitude changes as well as frequency changes. As can be gathered, the limiter is a form of clipping device, a circuit whose output tends to remain constant despite changes in the input signal. Most limiters behave in this fashion, provided that the input voltage remains within a certain range. The common type of limiter uses two separate electrical effects to provide a relatively constant output. These are leak-type bias and early (collector) saturation.

Operation of the amplitude limiter Figure 7-25 shows a typical FET amplitude limiter. Examination of the dc conditions shows that the drain supply voltage has been dropped through resistor R_D. Also, the bias on the gate is leak-type bias supplied by the parallel R_g–C_g combination. Finally, the FET is shown neutralized by means of capacitor C_N, in consideration of the high frequency of operation.

Leak-type bias provides limiting, as shown in Fig. 7-26. When input signal voltage rises, current flows in the R_g–C_g bias circuit, and a negative voltage is developed across the capacitor. It is seen that the bias on the FET is increased in proportion to the size of the input voltage. As a result, the gain of the amplifier is lowered, and the output voltage tends to remain constant.

Although some limiting is achieved by this process, it is insufficient by itself—the action just described would occur only with rather large input voltages. To overcome this, early saturation of the output current is used, achieved by means of a low drain supply voltage. This is the

[1] This does not include the ratio detector which (as is shown in Sec. 7-4.4) provides a fair amount of limiting.

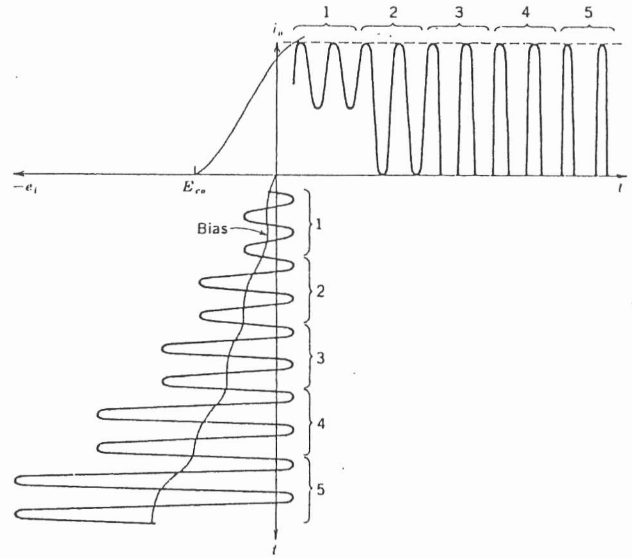


Fig. 7-26 Amplitude limiter transfer characteristic.

reason for the drain dropping resistor of Fig. 7-25. The supply voltage for a limiter is typically one-half of the normal dc drain voltage. The result of early saturation is to ensure limiting for conveniently low input voltages. However, it is possible for the gate-drain section to become forward-biased under saturation conditions, causing a short circuit between input and output. To avert this, a resistance of a few hundred ohms is placed between the drain and its tank. This is R of Fig. 7-25.

Figure 7-27 shows the response characteristic of the amplitude limiter. It indicates clearly that limiting takes place only for a certain range of input voltages, outside which output varies with input. Referring simultaneously to Fig. 7-26, we see that as input increases from value 1 to value 2, output current also rises. Thus no limiting has yet taken place. However, comparison of 2 and 3 shows that they both yield the same output current and voltage. Thus limiting has now begun. Value 2 is the point at which limiting starts, and is called the *threshold*

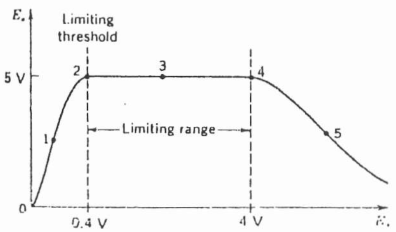


Fig. 7-27 Typical limiter response characteristic.

of limiting. As input increases from 3 to 4, there is no rise in output; all that happens is that the output current flows for a somewhat shorter portion of the input cycle. This, of course, suggests operation like that of a class C amplifier. Thus the *flywheel effect* of the output tank circuit is used here also, to ensure that the output voltage is sinusoidal, even though the output current flows in pulses. When the input voltage increases sufficiently, as in value 5, the angle of output current flow is reduced so much that less power is fed to the output tank. Therefore the output voltage is reduced. This happens here for all input voltages greater than 4, and this value marks the upper end of the limiting range, as shown in Fig. 7-27.

Performance of the amplitude limiter It has been shown that the range of input voltages, over which the amplitude limiter will operate satisfactorily, is itself limited. The limits are the threshold point at one end and the reduced angle of output current flow at the other end. In a typical practical limiter, the input voltage 2 may correspond to 0.4 V, and 4 may correspond to 4 V. The output will be about 5 V for both values and all voltages in between (note that all these voltages are peak-to-peak values). The practical limiter will therefore be fed a voltage which is normally in the middle of this range, that is, 2.2 V peak-to-peak or approximately 0.8 V rms. It will thus have a possible range of variation of 1.8 V (peak-to-peak) within which limiting will take place. In turn, this means that any spurious amplitude variations must be quite large compared to the signal to escape being limited.

Further limiting It is quite possible for the amplitude limiter described to be inadequate to its task, because signal-strength variations may easily take the average signal amplitude outside the limiting range. As a result, further limiting is required in a practical FM receiver.

Double limiter

This consists of two amplitude limiters in cascade, an arrangement that increases the limiting range very satisfactorily. Numerical values given to illustrate limiter performance showed an output voltage (all values peak-to-peak, as before) of 5 V for any input within the 0.4- to 4-V range, above which output gradually decreases. It is quite possible that an output of 0.6 V is not reached until the input to the first limiter is about 20 V. If the range of the second limiter is 0.6 to 6 V, it follows that all voltages between 0.4 and 20 V fed to the double limiter will be limited. This will be done by either one or both of the stages, and will yield constant output of 6 V. The use of the double limiter is thus seen to have increased the limiting range quite considerably.

AGC

A suitable alternative to the double limiter is automatic gain control. This is to ensure that the signal fed to the limiter is within its limiting range, regardless of the input signal strength, and also to prevent overloading of the last IF amplifier. If the limiter used has leak-type bias, then this bias voltage will vary in proportion to the input voltage (as shown in Fig. 7-26) and may therefore be used for AGC. If (as with a number of transistor limiters) leak-type bias is not used [4], a separate AGC detector is required. This stage takes part of the output of the last IF amplifier and rectifies and filters in the usual manner.

7-4.3 Basic FM Demodulators

The function of a frequency-to-amplitude changer, or FM demodulator, is to change the frequency deviation of the incoming carrier into an AF amplitude variation (identical to the one that originally caused the frequency variation). This conversion should be done efficiently and linearly. In addition, the detection circuit should (if at all possible) be insensitive to amplitude changes and should not be too critical in its adjustment and operation. Generally speaking, this type of circuit converts the frequency-modulated IF voltage of constant amplitude into a voltage that is both frequency- and amplitude-modulated. This latter voltage is then applied to a detector arrangement, which detects the amplitude change but ignores the frequency variations. It is now necessary to devise a circuit which has an output whose amplitude depends on the frequency deviation of the input voltage.

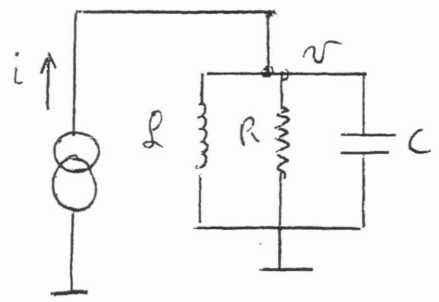
Slope detection Consider a frequency-modulated signal fed to a tuned circuit whose resonant frequency is to one side of the center frequency of the FM signal. The output of this tuned circuit will have an amplitude that depends on the frequency deviation of the input signal; this is illustrated in Fig. 7-28. As shown, the circuit is detuned by an amount δf, to bring the carrier center frequency to point A on the selectivity curve (note that A' would have done just as well). Frequency variation produces an output voltage proportional to the frequency deviation of the carrier, as shown.

This output voltage is applied to a diode detector with an RC load of suitable time constant. The circuit is, in fact, identical to that of an AM detector, except that the secondary winding of the IF transformer is off-tuned. (In a desperate emergency, it is possible, after a fashion, to receive FM with an AM receiver, with the simple expedient of giving the slug of the coil to which the detector is connected two turns clockwise. Remember to reverse the procedure after the emergency is over!)

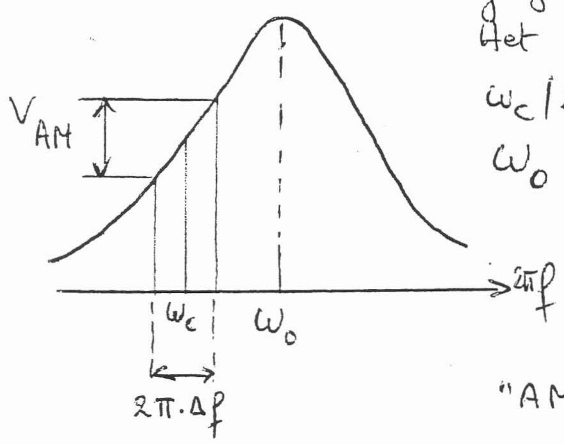
The slope detector does not really satisfy any of the conditions

1.2.2. F.M. demodulator

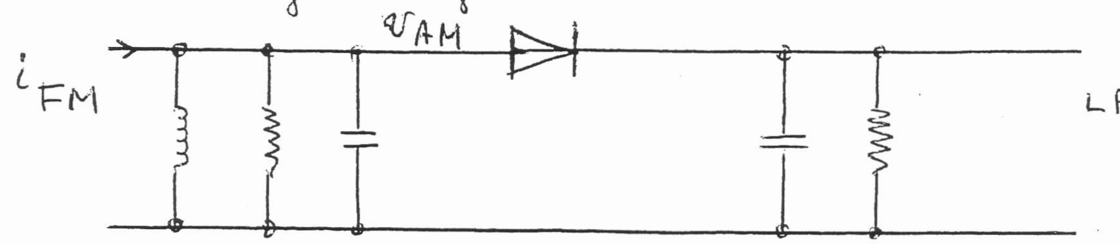
1.2.2.1. Met verstemde LC-kring



We beschouwen het hier naast gegeven schema waarbij een LC-parallelkring aangesloten is op de kollektorbak van een transistor. De karakteristiek van de LC kring is hieronder gegeven.



Het FM-signaal bezit een draaggolffrekwentie $\omega_c / 2\pi$ en men stemt het LC-filter af op $\omega_0 = \omega_{res} \neq \omega_c$, zodat ω_c in het lineaire gebied van de zijflank ligt. In dit geval ontstaat t.g.v. de FM-golf een spanning v die "AM" is, immers frekwentie wijzigingen rond ω_c worden omgezet in veranderingen van amplitude. De principiële FM-demodulator is dus van de volgende gedaante:



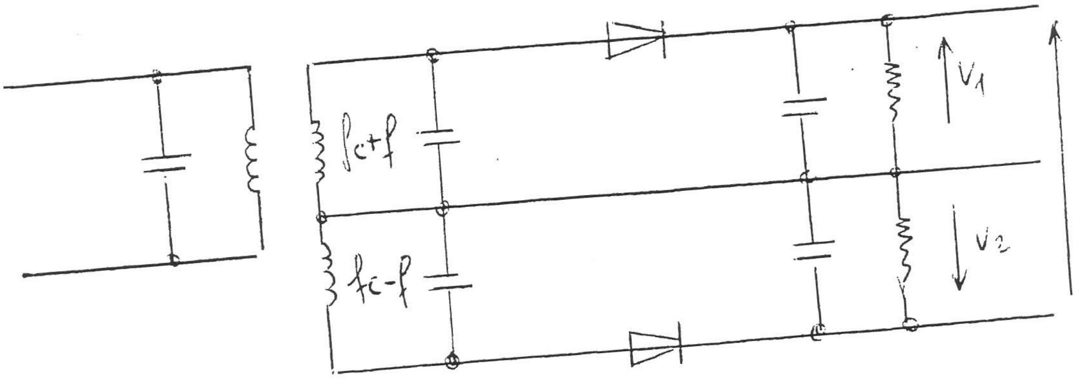
Nadelen van deze schakeling zijn:

- geringe gevoeligheid
- gering lineair werkingsgebied.

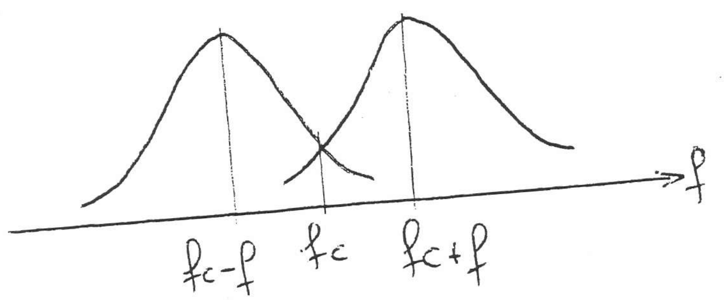
1.2.2.2. Met 2 verstemde LC-kringen

Om het lineair werkingsgebied te vergroten kan men werken met 2 verstemde LC-kringen, afgeregeld op $f_c + f$ en $f_c - f$, terwijl het signaal, d.m.v. een trafokoppeling van de primaire overgebracht wordt naar 2 sekundaire wikkelingen zodat de signalen hier in tegenfaze zijn t.o.v. het fiktieve gemeenschappelijke aardpunt.

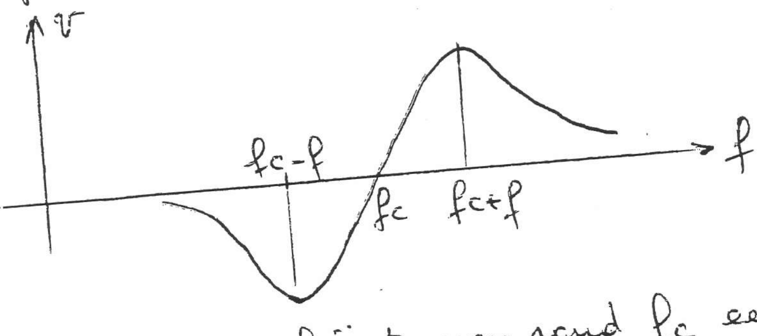
Het schema is op de volgende bladzijde gegeven:



Het verloop van V_1 en V_2 is uit onderstaande figuur af te leiden:

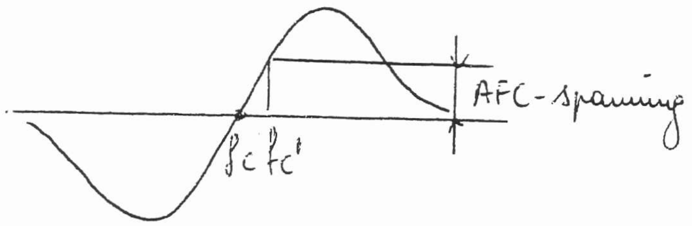


Daar de uitgangsspanning $v = V_1 - V_2$ heeft men volgend spanningsverloop in functie van de frekwentie:



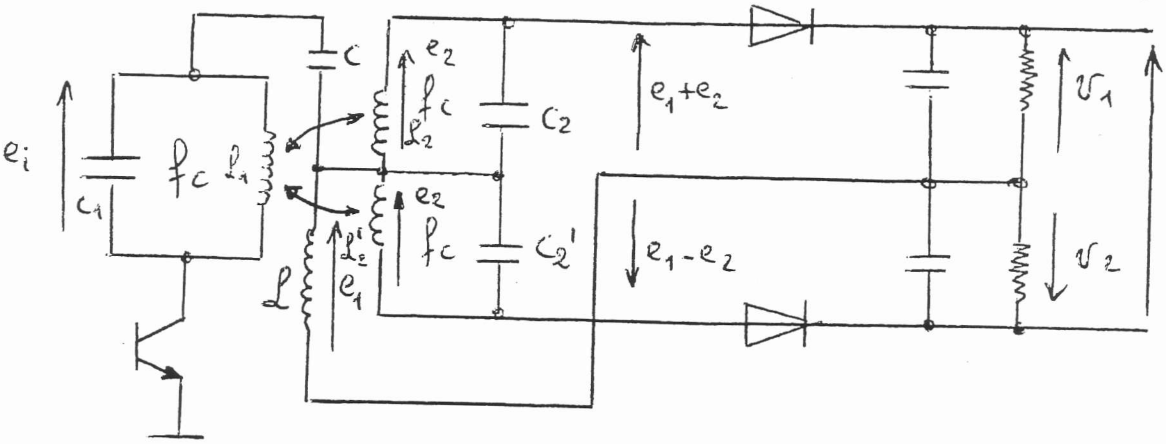
Op deze wijze verkrijgt men rond f_c een breed lineair werkingsgebied. De diodes zorgen er dus voor dat telkens de omhullende gedetekteerd wordt (de positieve omhullende t.o.v. de fiktieve massa) zodat inderdaad bovenstaand spanningsverloop geldig is.
Voordeel van de methode is dat indien het signaal niet gemoduleerd is, en de zenderfrekwentie juist ingesteld is, V precies nul is. Als f_c verandert omdat de lokale oscillator verloopt, geeft dit meteen aanleiding tot een gemiddelde DC spanning van V, die een maat is voor de afwijking en de zin van de afwijking. Op die manier kan

men eenvoudig AFC (automatic frequency control) toepassen door deze DC spanning te sturen naar de lokale oscillator, uitgevoerd als VCO.

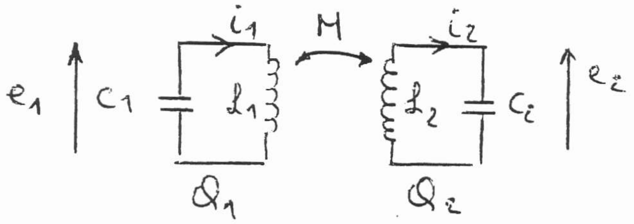


1.2.2.3. Foster-Seely detektor

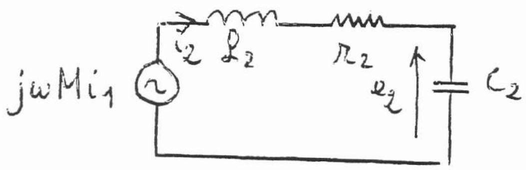
Hier wordt er gewerkt met dubbel afgestemde kringen die zwak met elkaar gekoppeld zijn. Het schema is hieronder gegeven:



De koppelkondensator c is groot genomen, zodat er geen faseverschuiving over optreedt, zodat e_i en e_1 in fase zijn.
De spoel L is een RF-smoorspoel zodat $e_i = e_1$
Beschouwen we de dubbel afgestemde kringen $C_1 L_1$ en $C_2 L_2$:



We kunnen de rechter afgestemde kring vervangen door volgend ekwivalent:



$j\omega M i_1$ is de geïnduceerde spanning in de rechterspoel t.g.v. i_1 zodat

$$i_1 s M = i_2 \left(r_2 + s L_2 + \frac{1}{sC_2}\right)$$

Daar $i_1 = e_1 s C_1$ en $i_2 = e_2 s C_2$ komt er:

$$\frac{e_2}{e_1} = \frac{s C_1 M}{C_2 \left(r_2 + s L_2 + \frac{1}{s C_2} \right)} \qquad (1)$$

Bij resonantie, t.t.z. $f_c = \frac{1}{2\pi\sqrt{L_2 C_2}}$ geldt er

$$\frac{e_{2c}}{e_1} = \frac{s_c C_1 M}{C_2 r_2} = j\omega_c \frac{C_1 M}{C_2 r_2} \qquad (2)$$

Deling van (1) door (2) en vervanging van s door $j\omega$ geeft:

$$\frac{e_2}{e_{2c}} = \frac{\omega}{\omega_c} \cdot \frac{1}{1 + j\omega \frac{L_2}{r_2} + \frac{1}{j\omega C_2 r_2}}$$

Benaderend mag men nu $\omega \simeq \omega_c$ stellen, daar men toch dicht bij resonantie werkt

(FM omroep: I.F. $= f_c = 10{,}7$ MHz en $\Delta f = 75$ kHz)

$$\frac{e_2}{e_{2c}} = \frac{1}{1 + j\omega \frac{L_2}{r_2} + \frac{1}{j\omega C_2 r_2}}$$

$$= \frac{1}{1 + j\left[\frac{\omega_c}{\omega_c} \frac{\omega L_2}{r_2} - \frac{\omega_c}{\omega \omega_c C_2 r_2} \right]}$$

Met $Q_2 = \frac{\omega_c L_2}{r_2} = \frac{1}{\omega_c C_2 r_2}$ komt er nog:

$$\frac{e_2}{e_{2c}} = \frac{1}{1 + j Q_2 \left(\frac{\omega}{\omega_c} - \frac{\omega_c}{\omega} \right)}$$

Hierbij is $\frac{\omega}{\omega_c} - \frac{\omega_c}{\omega} = \frac{(\omega - \omega_c)(\omega + \omega_c)}{\omega_c \omega} \simeq \frac{\Delta\omega\, 2\omega_c}{\omega_c^2} = \frac{2\Delta\omega}{\omega_c} = \alpha = \frac{2\Delta f}{f_c}$

zodat

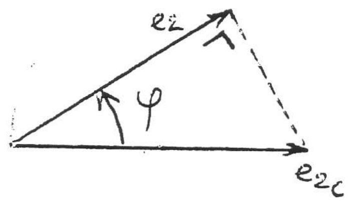
$$\frac{e_2}{e_{2c}} = \frac{1}{1 + j\alpha Q_2}$$

$$\measuredangle \frac{e_2}{e_{2c}} = -\mathrm{Bgtg}(\alpha Q_2) \quad \text{of} \quad -\mathrm{tg}\,\varphi = \alpha Q_2$$

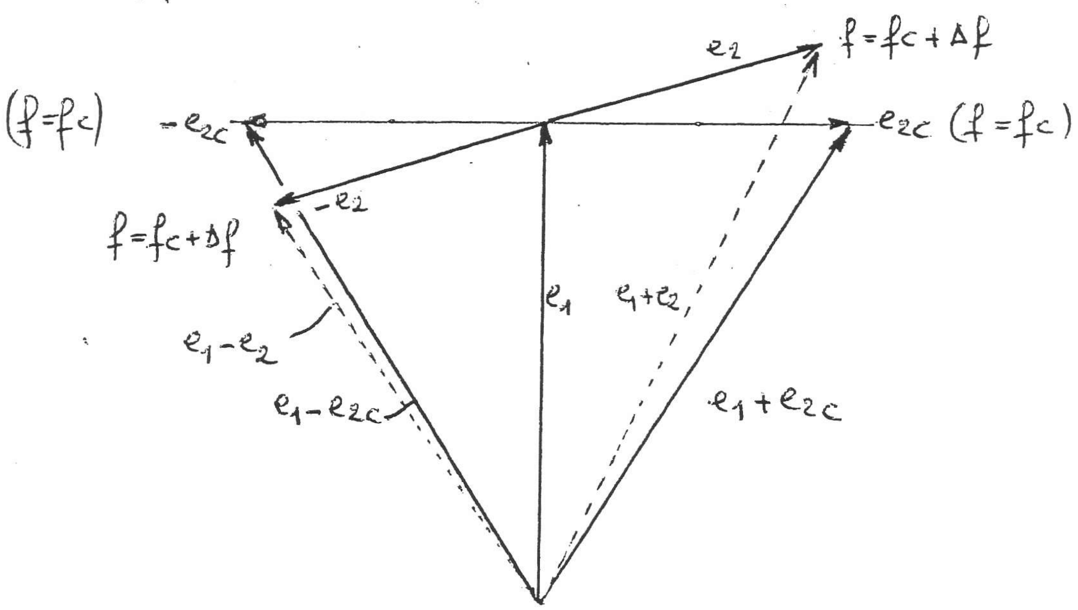
zodat

$$\left\| \frac{e_2}{e_{2c}} \right\| = \frac{1}{\sqrt{1 + \mathrm{tg}^2 \varphi}} = \cos \varphi$$

Dit betekent dat $\left\|\frac{e_2}{e_{2c}}\right\|$ op een cirkel voor te stellen is.
e_{2c} is de diameter, terwijl e_2 een punt van een cirkel voorstelt zoals in onderstaande figuur:



Wanneer φ nul is, is $e_2 = e_{2c}$ en dan is $f = f_c$
Voor de onderste afgestemde kring kan men een analoge vergelijking opstellen ($C_2' R_2'$ t.o.v. e_1)
Daar we nu echter V aftakken als $v_1 - v_2$, moeten we de 2e cirkel naast de eerste tekenen.
Verder heeft men dat t.o.v. het fiktieve massapunt e_1 er moet bijgeteld worden, waarbij we 90° fase-verschuiving van e_1 in rekening moeten brengen, inderdaad men bewijst dat bij afgestemde kringen er tussen e_i en e_2 een faseverschil van 90° bestaat bij f_c, zodat we met bovenstaande beschouwingen tot volgend vektordiagram komen:



Bij de centrale frekwentie $f = f_c$ wordt $V = v_1 - v_2$, met $v_1 = \|e_1 + e_{2c}\|$ en $v_2 = \|e_1 - e_{2c}\|$ gelijk aan nul.
Als f verandert met een bedrag Δf verschuift $e_1 + e_2$ in de ene richting (bvb. verlenging), terwijl $e_1 - e_2$ verschuift

in de andere richting (verkorting). Dit impliceert dat de amplitude $V = v_1 - v_2 > 0$

Als nu φ klein is zijn de veranderingen lineair te beschouwen wat betekent dat αQ klein moet genomen worden.

Stel dat voor $\varphi_{max} = 20°$ genomen wordt, dan moet

$$\frac{2\Delta f}{f_c} Q_2 < \mathrm{tg}\, 20$$

Met $f_c = 10{,}7$ MHz en $\Delta f = 75$ kHz heeft men

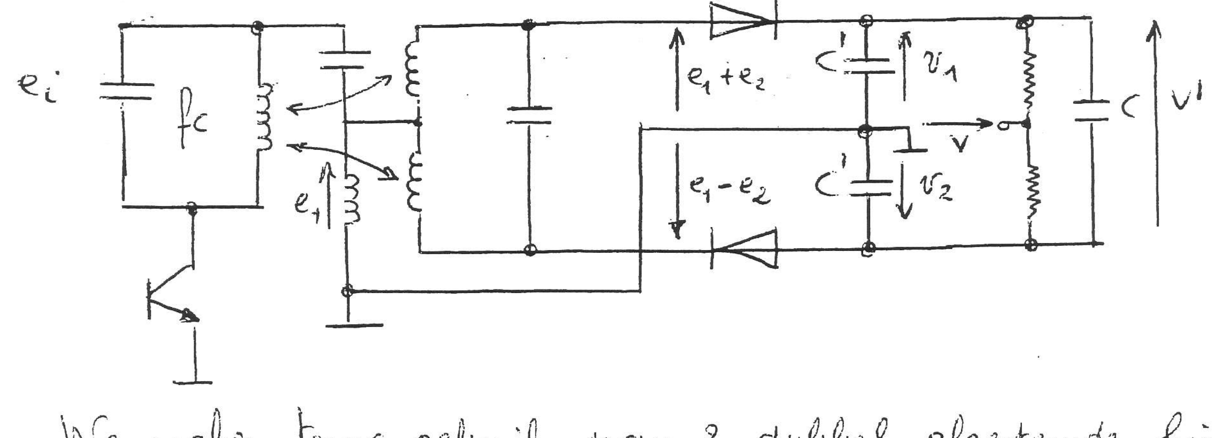
$$Q_{2max} = 26.$$

Zelfs 20° geeft in de praktijk te sterke niet lineariteiten, vandaar dat Q opzettelijk laag genomen wordt (dus "slechte" afgestemde kringen).

Men heeft terug de mogelijkheid om AFC in te bouwen.

Als enig nadeel van de Foster - Seely detektor noteert men het feit dat amplitude variaties van het ingangssignaal in het uitgangssignaal terecht komen. De detektor is dus "AM-gevoelig". De volgende detektor vertoont dit nadeel niet.

1.2.2.4. Ratio detektor (verhoudingsdetektor)



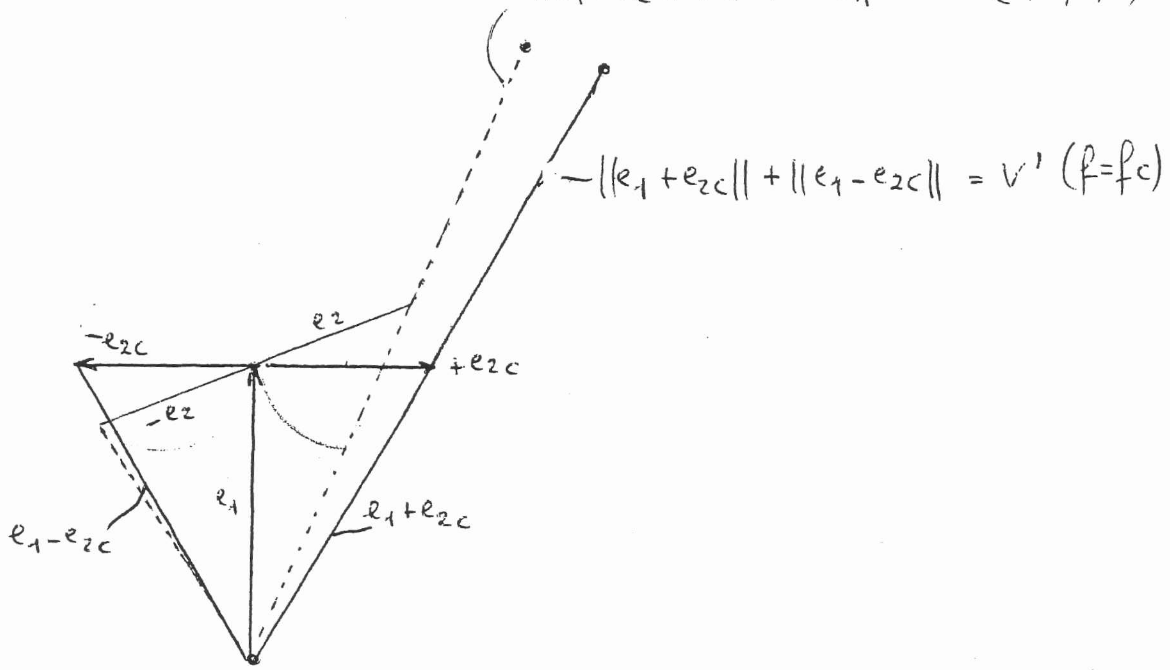
We maken terug gebruik van 2 dubbel afgestemde kringen.
Aan de ingang van de 2 detektoren is niets gewijzigd, $e_1 + e_2$ en $e_1 - e_2$ worden aangelegd.
Daar de onderste diode omgekeerd werd verkrijgen we nu:

$$v_1 = \|e_1 + e_2\|$$
$$v_2 = -\|e_1 - e_2\|$$

zodat $V' = \|e_1 + e_2\| + \|e_1 - e_2\|$

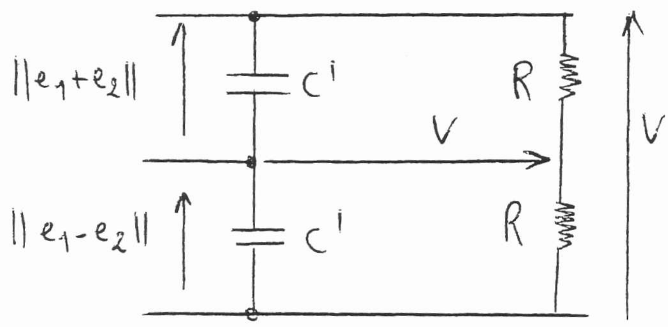
Wanneer we vorig vektordiagram nemen, dan krijgen we voor V' een konstante spanning onafhankelijk van f (of toch bijna)

$$\|e_1+e_2\| + \|e_1-e_2\| = v' \quad (f \neq f_c)$$



Wanneer we C groot kiezen zal v' niet ogenblikkelijk veranderen zodat AM-variaties verdwijnen. Op lange termijn is V' een maat voor de amplitude van e_i of dus voor de gemiddelde sterkte van het ontvangstsignaal
v' kan dus gebruikt worden als AGC (of AVC) spanning om de versterking van de H.F.-versterkertrappen aan te passen zodat het audioniveau konstant is.

De eigenlijke uitgangsspanning v kan nu uit volgend schema afgeleid worden:



De uitgangsspanning wordt

$$V = \frac{V'}{2} - \|e_1 - e_2\|$$

$$= \frac{\|e_1+e_2\|+\|e_1-e_2\|}{2} - \|e_1-e_2\|$$

$$= \frac{\|e_1+e_2\| - \|e_1-e_2\|}{2}$$

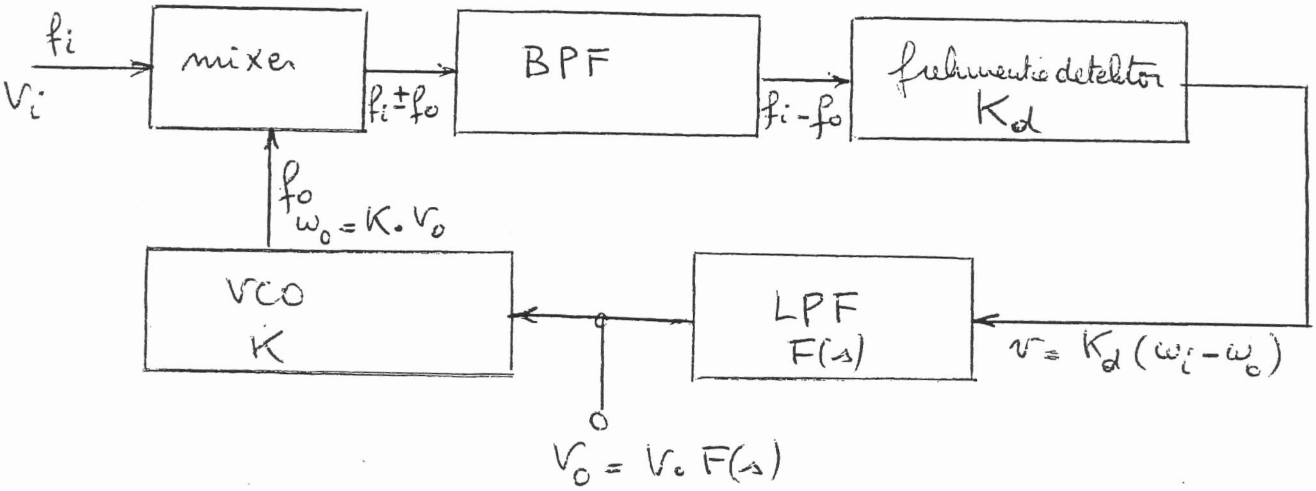
Deze spanning is dus op een faktor 1/2 na de uitgangsspanning van de Foster-Seely-detektor met goede eigenschappen:

- lineair
- AFC-regeling mogelijk

met bijkomstige mogelijkheid van AGC-regeling.
Ze worden gebruikt tot 40 MHz (IF van TV)
Voor de FM-omroep (87,5–108 MHz) vormt dit geen probleem daar er aan frekwentiekonversie gedaan wordt, waardoor de detektie bij de middenfrekwentie (IF = 10,7 MHz) gebeurt.

1.2.2.5. Frequency following (FF) of frequency compressive feedback (FCF)-detektor

Men heeft hier te maken met een regelsysteem waarbij negatieve terugkoppeling wordt toegepast.



De mixer mengt de signalen of frekwentie f_i en f_o (afkomstig van de VCO) zodat som en verschilfrekwenties ontstaan.
Hierbij wordt $f_i > f_o$ genomen. De BPF maakt dat de verschilfrekwentie $f_i - f_o$ overblijft.
De frekwentiedetektor geeft een spanning v af die evenredig is met het frekwentieverschil:

$$v = K_d(\omega_i - \omega_o)$$

De frekwentiedetektor is een der vorige FM-demodulatoren
$F(s)$ is een klassiek LPF waarvan de keuze afhangt van het gewenste werkingsgebied, ruis en stabiliteit.
We stellen nu de transferfunktie op, waarbij de BPF buiten beschouwing gelaten wordt. De taak van de BPF is

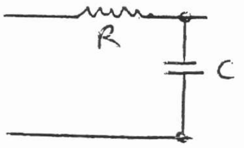
immers enkel de verschilfrekwentie doorlaten.
Dit impliceert dat we ook faseverschuivingen van de BPF zomaar nul stellen (in de praktijk kan dit dus niet om stabiliteitsredenen).

$$K_d(\omega_i - \omega_o)\, F(s) = V_o$$

of nog $K_d(\omega_i - K.V_o) F(s) = V_o$
waaruit volgt:

$$\frac{V_o}{\omega_i} = \frac{K_d\, F(s)}{1 + K_d . K . F(s)}$$

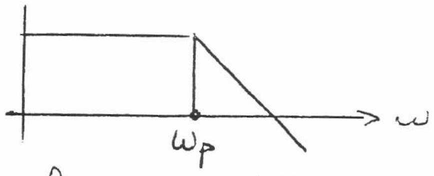
Kiest men nu als LPF een 1e orde RC filter met



$$F(s) = \frac{1}{sRC + 1}$$

dan komt er:

$$\frac{V_o(s)}{\omega_i(s)} = \frac{K_d}{sRC + (1 + K_d \cdot K)}$$



dit in de onderstelling dat de andere komponenten frekwentieonafhankelijk zijn.
met $\omega_p = \frac{1 + K_d . K}{RC}$

Hieruit leiden we af dat de transferfunktie V_o in funktie van de ogenblikkelijke frekwentie ω_i geeft. Dit is een eerste orde filter zelf, waardoor snelle frekwentieveranderingen van het ingangssignaal (op ω_i) niet gevolgd worden wanneer ze hoger dan ω_p zijn.
Voorbeelden van snelle variaties zijn o.a. ruis.
Zelfs onder relatief slechte signaal-ruisverhoudingen van het HF-signaal zal deze detektor nog relatief goed werken.
Men kan dit ook inzien door het systeem te beschouwen met tijdsignalen:
De ingangsfrekwentie $f_i = f_c + \Delta f . e(t)$ met f_i de ogenblikkelijke frekwentie en f_c de draaggolffrekwentie.

De frekwentie die aan de ingang van de frekwentiedetektor gelegd wordt is dus

$$f_i - f_o = (f_c + \Delta f \cdot e(t)) - K \cdot V_o$$

met $K \cdot V_o = f_o = \dfrac{K K_d \cdot \omega_i}{1 + K \cdot K_d}$

$$= \frac{K \cdot K_d \, (f_c + \Delta f e(t))}{1 + K \cdot K_d}$$

Dus wordt

$$f_i - f_o = \frac{f_c + \Delta f \, e(t)}{1 + K \cdot K_d}$$

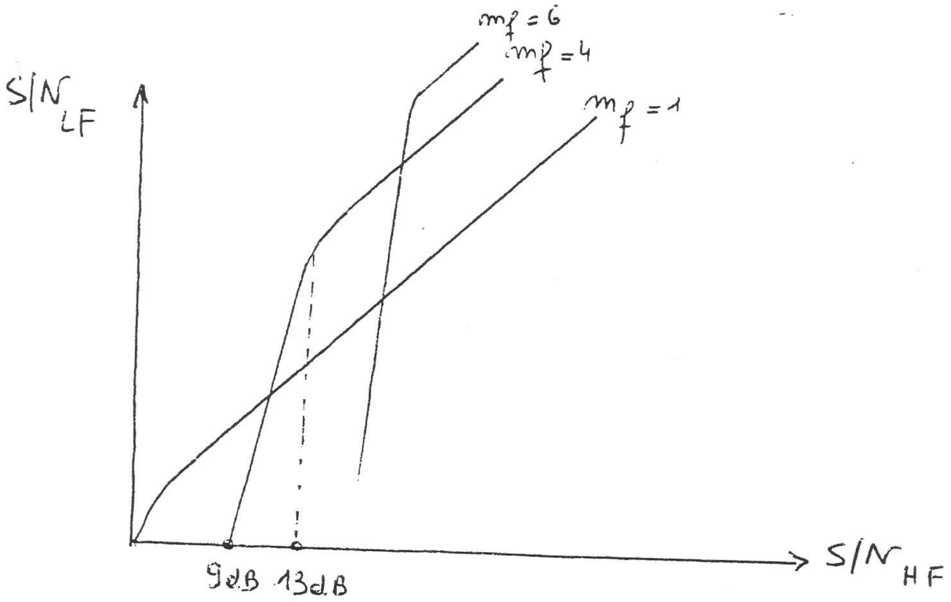
Het komt er dus op neer dat de frekwentiedetektor een FM-signaal te detekteren krijgt met effektieve frekwentiezwaai

$$\Delta f_{eff} = \frac{\Delta f}{1 + K \cdot K_d}$$

Δf wordt dus "ineengedrukt" (compressive), waardoor meestal overgegaan wordt van breedband naar smalband FM. Normaal gezien is de signaal-ruis verhouding bij FM aan de uitgang evenredig met de signaal-ruisverhouding van de drager:

$$S/N_{uit} = 3 m_f^2 \cdot S/N_{drager}.$$

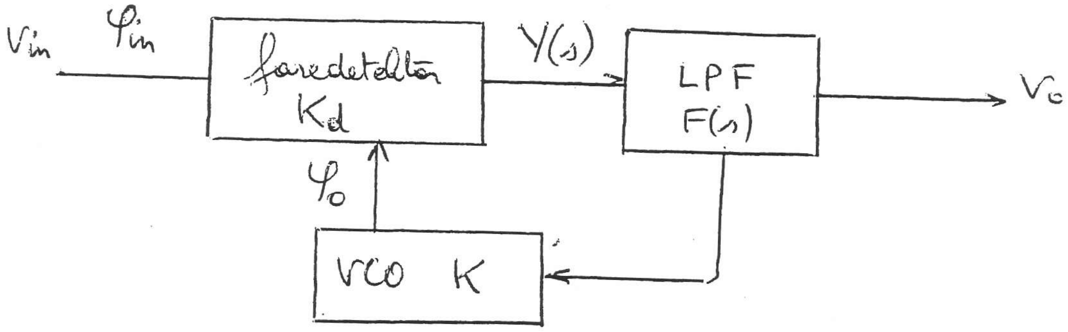
Bij FM treedt echter het zgn <u>treshold-effekt</u> op: Wanneer S/N drager zeer slecht wordt (dus zeer klein), wat veel ruis impliceert, dan heeft dit een plotse verslechtering van de S/N_{uit} (= LF) tot gevolg, zodat bovenstaande formule niet meer opgaat. Deze eigenschap is enkel <u>geldig bij breedband FM</u>, en doet zich voor bij of beter beneden hogere waarden van S/N drager bij stijgende m_f.
Dit betekent dat bij slechte S/N drager smalband FM (kleine m_f) toch betere resultaten zal geven.
Vandaar dat bovenstaande detektor goede resultaten geeft bij slechte signaal-ruisverhoudingen van het H.F.-ontvangstsignaal.
Op de volgende blz. is $S/N_{LF} = f(S/N_{HF}, m_f)$ gegeven.



← | treshold bij mf = 4.

1.2.2.6. PLL-demodulator

Deze detektor heeft een nog beter ruisgedrag dan de vorige daar hier fazen vergeleken worden. Bovendien zijn minder komponenten vereist.



De fazedetektor is meestal opgebouwd rond een vermenigvuldiger, immers $\cos(\omega t + \phi_1)\cos(\omega t + \phi_2) = \frac{1}{2}\cos(\phi_1 - \phi_2) + \frac{1}{2}\cos(2\omega t + \phi_1 + \phi_2)$

De trilling op $2\omega t$ wordt weggefilterd.

Hier heeft men dus aan de uitgang van de fazedetektor:

$$Y(s) = K_d(\varphi_{in} - \varphi_o)$$

en voor de spanningsgestuurde oscillator (VCO)

$$\omega = K \cdot V_o$$

met $\omega = \dfrac{d\varphi_o}{dt}$

of nog in het s-domein:

$$V_o(s) = \frac{s\varphi_o}{K}$$

Verder geldt:

$$V_0(s) = F(s) \cdot K_d \left[\varphi_{in}(s) - \frac{K \cdot V_0(s)}{s} \right]$$

waaruit volgt:

$$V_0(s) = \varphi_{in}(s) \cdot \frac{s \cdot F(s) K_d}{s + K_d \cdot K \cdot F(s)}$$

of in termen van de ingangsfrekwentie $\omega_i(s)$:

$$V_0(s) = \omega_i(s) \frac{F(s) \cdot K_d}{s + K_d \cdot K \cdot F(s)}$$

V_0 volgt dus terug de veranderingen van de ingangsfrekwentie ω_i, waarbij echter de transferfunktie in rekening dient gebracht.
Kiezen we bvb. terug een 1e orde LPF (RC-type), met

$$F(s) = \frac{1}{RCs + 1}$$

dan komt er:

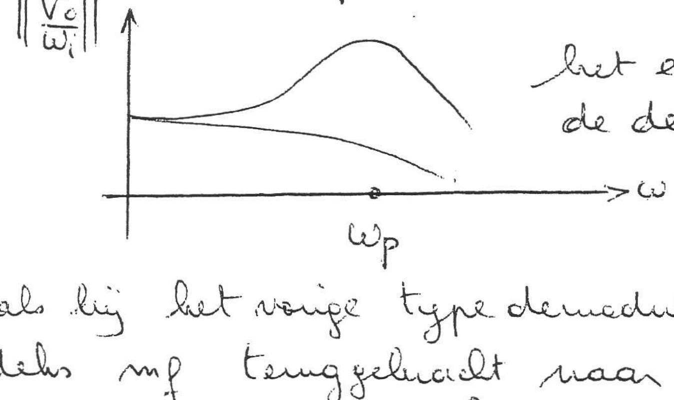
$$V_0(s) = \omega_i(s) \cdot \frac{K_d}{s^2 RC + s + K \cdot K_d}$$

of in de standaardvorm:

$$\frac{V_0(s)}{\omega_i(s)} = \frac{K_d / RC}{s^2 + \frac{s}{RC} + \frac{K K_d}{RC}}$$

met resonantiepulsatie

$$\omega_p = \sqrt{\frac{K \cdot K_d}{RC}}$$



het exakte verloop hangt af van de demping.

Zoals bij het vorige type demodulator wordt de modulatie-index m_f teruggebracht naar

$$m_{f_{eff}} = \frac{m_f}{1 + K \cdot K_d}$$

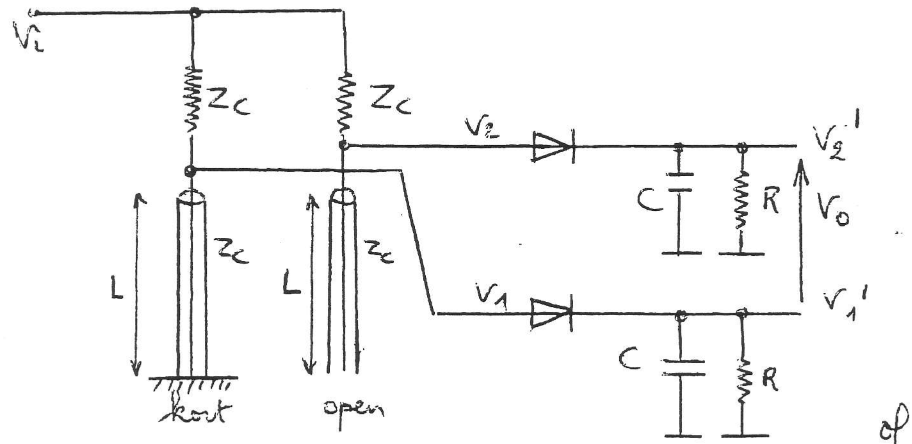
Door het tresholdeffekt wordt terug de signaal-ruisverhouding bij slechte HF-signalen verbeterd
We merken nog op dat inwendig (zonder F(s)) reeds

automatisch een 1e orde LPF- ingebouwd is, wat tot gevolg heeft dat de orde van het volledige systeem (inklusief F(s)) één hoger is dan F(s) zelf.
Het enige soort onvoorwaardelijkstabiele PLL-systemen mag dus slechts een 1e orde LPF als F(s) benutten. Is F(s) van hogere orde dan kan het PLL-systeem onstabiel worden.

PLL-systemen worden meestal geïntegreerd. Ze hebben een werkingsgebied dat gaat tot 30 MHz.
voorbeelden zijn de MC 4044 met VCO MC 4024, verder heeft men de NE560 en de LM 565 en de CD4046.

1.2.2.7. FM-demodulator met transmissielijnen

We beschouwen volgend transmissielijnsysteem, met karakteristieke impedanties Zc. De ene transmissielijn is open, terwijl de andere kortgesloten werd.



De ingangsimpedantie vd kortgesloten lijn is

$Z_{in_{kort}} = j\, Z_c\, tg\, kL$

zodat

$$V_1 = \frac{j\, tg(kL)}{1 + j\, tg(kL)}\, V_{in}$$

of $V_1 = j \cdot e^{-jkL} \sin(kL) \cdot V_{in}$

De ingangsimpedantie van de open lijn is

$Z_{in_{open}} = -j Z_c \cotg(kL)$

zodat

$$V_2 = \frac{-j \cotg(kL)}{1 - j \cotg(kL)}\, V_{in} = j e^{+jkL} \cos(kL)\, V_{in}$$

Detektie van de amplitudes van V_1 en V_2 geeft

$v_1' = \sin(kL)$ en $v_2' = \cos(kL)$

Neemt men de uitgang of tussen V_2' en V_1' dan komt er

$V_0 = V_2' - V_1' = V_{in}(\cos(kL) - \sin(kL)) = V_{in}\sqrt{2}\ \cos\left(kL + \frac{\pi}{4}\right).$

Neemt men nu als lengte van de 2 lijnen $L = \lambda_0/8$, dan heeft men:

$$V_0 = \sqrt{2}\ \cos\left(\frac{\pi}{2}\right) = 0$$

dit wanneer de golflengte van het inkomend signaal gelijk is aan de golflengte waarvoor L berekend werd.

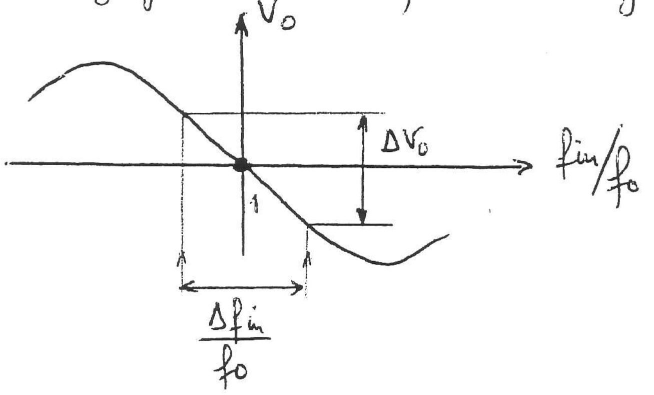
Als $\lambda_{in} < \lambda_0$ of als $f_{in} > f_0$ dan is

$$V_0 = \sqrt{2}\ \cos\left(\frac{\pi}{4}\cdot\frac{\lambda_0}{\lambda_{in}}\right) = \sqrt{2}\cos\left(\frac{\pi}{4}\ \frac{f_{in}}{f_0}\right) < 0$$

Als $\lambda_{in} > \lambda_0$ of als $f_{in} < f_0$ dan is

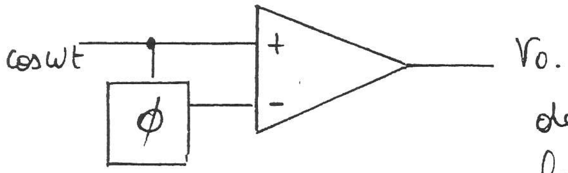
$$V_0 = \sqrt{2}\ \cos\left(\frac{\pi}{4}\ \frac{f_{in}}{f_0}\right) > 0$$

Dit geeft aanleiding tot volgende karakteristiek:



Wanneer het signaal niet gemoduleerd is, is V_0 dus nul. Wijzigingen rond de centrale frekwentie f_0 geven dus aanleiding tot spanningsveranderingen aan de uitgang.

1.2.2.8. FM - demodulator met verschilversterker.



Het signaal aan de - ingang van de verschilversterker is $\cos(\omega t + \phi)$ t.g.v. het fazeverschuivend netwerk, zodat het uitgangssignaal evenredig is met

$$V_0 \div \cos\omega t - \cos(\omega t + \phi)$$

waarbij ω de ogenblikkelijke frekwentie van het FM-signaal is. Bovenstaande uitdrukking is te schrijven als

$$\cos\omega t - \cos\omega t\cos\phi + \sin\omega t \sin\phi$$

of

$$\cos\omega t\,(1 - \cos\phi) + \sin\omega t\ \sin\phi \equiv A\cos(\omega t + \theta)$$

Deze laatste vergelijking gaat op daar

$$A\cos(\omega t + \theta) = A\cos\omega t\cos\theta - A\sin\omega t\sin\theta$$

In dit geval moet gelden dat

$$1 - \cos\phi = A \cos\theta$$
$$\sin\phi = -A \sin\theta$$

Kwadrateren en optellen van deze 2 vergelijkingen geeft:

$$(1-\cos\phi)^2 + \sin^2\phi = A^2$$

waaruit volgt: $A = \sqrt{2}\sqrt{1-\cos\phi}$

Deling van de vergelijkingen geeft:

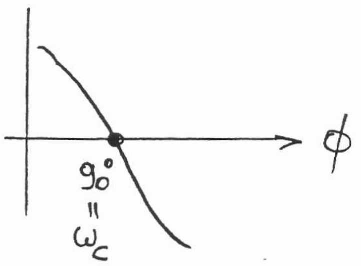
$$\operatorname{tg}\theta = \frac{-\sin\phi}{1-\cos\phi}$$

Dit betekent dat het uitgangssignaal kan geschreven worden als $V_0 = \sqrt{2}\sqrt{1-\cos\phi}\cos(\omega t+\theta)$

Als men nu de <u>amplitude</u> van V_0 detekteert, dan verkrijgt men een spanning die mee verandert met de faseveranderingen die het inkomend signaal ondergaat in de faseverschuiver ϕ.

Voor waarden van $\cos\phi$ tot rond 0,4 geldt benaderend dat $\sqrt{1-\cos\phi} \simeq 1 - \frac{1}{2}\cos\phi$

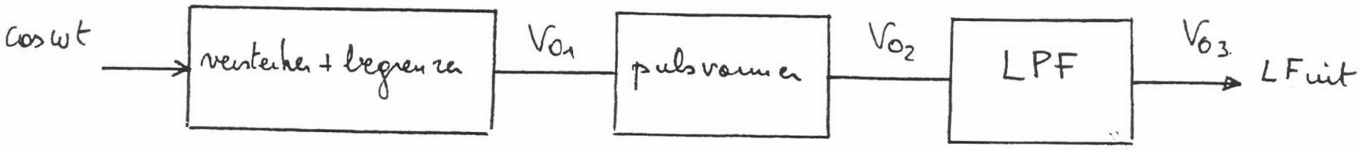
Voor $\phi = 90°$ verkrijgt men een niet gemoduleerde AM-draaggolf.



Kiest men dus voor de faseverschuiver een RLC-kring, dan heeft men op resonantiefrekwentie precies 90° faseverschuiving. Wijkt de frekwentie van het ingangssignaal of van de resonantiefrekwentie van de kring, dan wordt dit via een faseverandering omgezet in een amplitudeverandering van V_0.

Als men er nu voor zorgt dat de faseveranderingen gering zijn, dan werkt de detektor lineair.

1.2.2.9. Teldetektor.



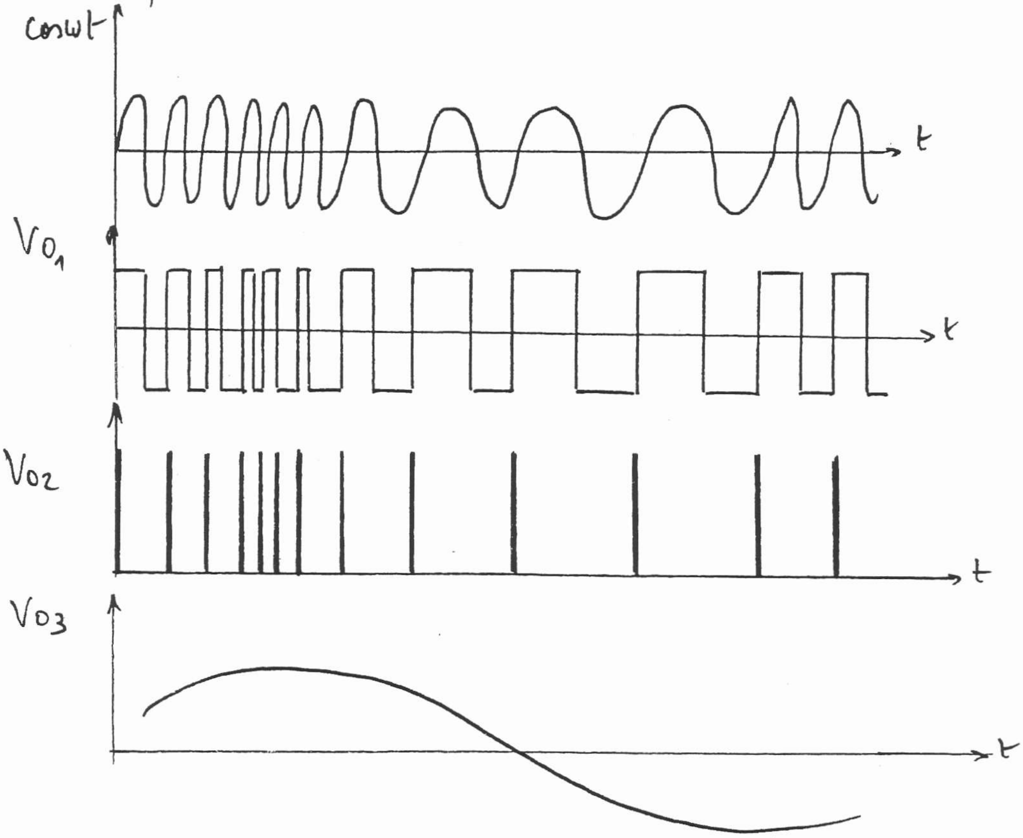
Het inkomend signaal wordt zeer sterk versterkt en (uiteraard) begrensd, zodat het sinusoïdaal signaal een blokgolf wordt. Iedere opgaande flank triggert een impulsvormer (= monostabiele multivibrator), zodanig dat pulspositiemodulatie verkregen wordt (PPM).

De detektie van het LF-signaal kan nu gebeuren door het signaal door een laagdoorlaatfilter te sturen.

Uiteraard moeten de gegenereerde impulsen van de monostabiele multivibrator korter zijn dan de kleinste periode van de HF-trilling om geen overlapping te verkrijgen.

Dit vergt (meestal) ECL-schakelingen.

De golfvormen zijn hieronder gegeven.



$$\bar{\nabla} \times \bar{E} = -j\omega\mu \bar{H} \qquad \bar{a} \times (\bar{b} \times \bar{c}) = (\bar{a}\bar{c})\bar{b} - (\bar{a}\bar{b})\bar{c}$$

$$\bar{\nabla} \times \bar{H} = +j\omega\varepsilon \bar{E}$$

$$\bar{\nabla} \cdot \bar{E} = 0$$

$$\bar{\nabla} \cdot \bar{H} = 0$$

$$Z_{in}' = \frac{Z_L' + j \, tg\, kL}{1 + j Z_L' \, tg\, kL} \qquad K(x) = K e^{-2jk(L-x)}$$

$$Z_W = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} \qquad k_c = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}$$

$$W_r = \iint_{\phi\,\theta} P_\Omega(\theta,\phi) \sin\theta \, d\theta \, d\phi \qquad A_{eff} = \frac{\lambda^2 G}{4\pi}$$

$$\bar{H}_\phi = j \frac{e^{-jkr} k I_m L \sin\theta}{4\pi r} \left(1 + \frac{1}{jkr}\right), \quad \bar{H}_r = \bar{H}_\theta = 0$$

$$\bar{E}_\theta = j \frac{e^{-jkr} k I_m L Z_0 \sin\theta}{4\pi r} \left(1 + \frac{1}{jkr} - \frac{1}{k^2 r^2}\right)$$

$$\bar{E}_r = \frac{e^{-jkr} I_m L Z_0 \cos\theta}{2\pi r^2} \left(1 + \frac{1}{jkr}\right)$$

$$\bar{E}_\theta = \frac{j Z_0 I_m e^{-jkr}}{2\pi r} \cdot \left(\frac{\cos\left(\frac{kL}{2} \cdot \cos\theta\right) - \cos\left(\frac{kL}{2}\right)}{\sin\theta}\right)$$

$$G = 6 \left(\frac{D}{\lambda}\right)^2 \qquad \phi_{3dB} = \left(170 \frac{\lambda}{D}\right)^\circ$$

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