An Extremely Low-Noise 96 MHz Crystal Oscillator for UHF/SHF Applications

Summarized from a Lecture at the VHF-UHF Convention 1980 in Weinheim, West Germany

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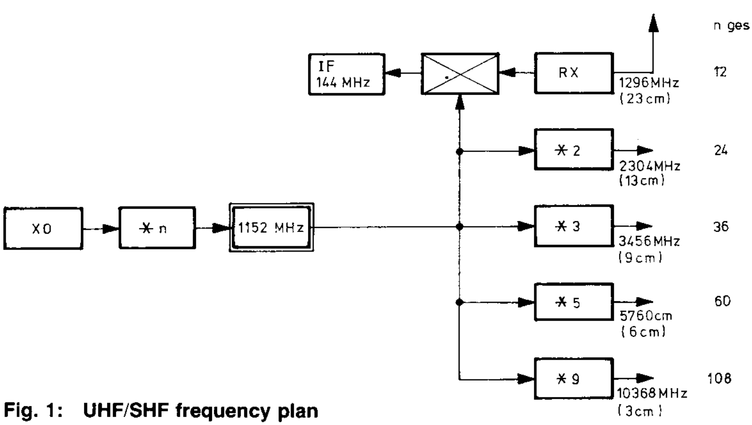
1. IMPORTANCE OF SHORT-TERM STABILITY AND PHASE NOISE IN TRANSMITTER AND RECEIVER APPLICATIONS

The output frequency of UHF and SHF equipment is to be obtained from a stable crystal oscillator with the aid of frequency multiplier chains.

Figure 1 shows the various frequency plans that can be used. All bands up to 3 cm can be obtained from a basic frequency of 1152 MHz. This is obtained with the aid of frequency doublers, and triplers from the given frequencies. Since it is difficult to manufacture crystal oscillators operating at a frequency of 192 MHz, and because 72 MHz and 144 MHz will cause unwanted spurious signals in the 2 m and 70 cm band (converter), the most suitable frequency seems to be 96 MHz.

The resulting total multiplication factor n_{tot} is given in the illustration, and is between 12 (23 cm) and 108 (3 cm). This results in very high demands on the frequency stability of the master oscillator. Whereas insufficient long-term stability will be noticeable as drift that can be avoided using a stable crystal oscillator circuit in a crystal oven and using aged crystals of sufficiently high Q, short-term stability is far more critical.



f_{XO}	n
72 MHz	16
96 MHz	12
128 MHz	9
144 MHz	8
192 MHz	6

Fig. 1: **UHF/SHF frequency plan**

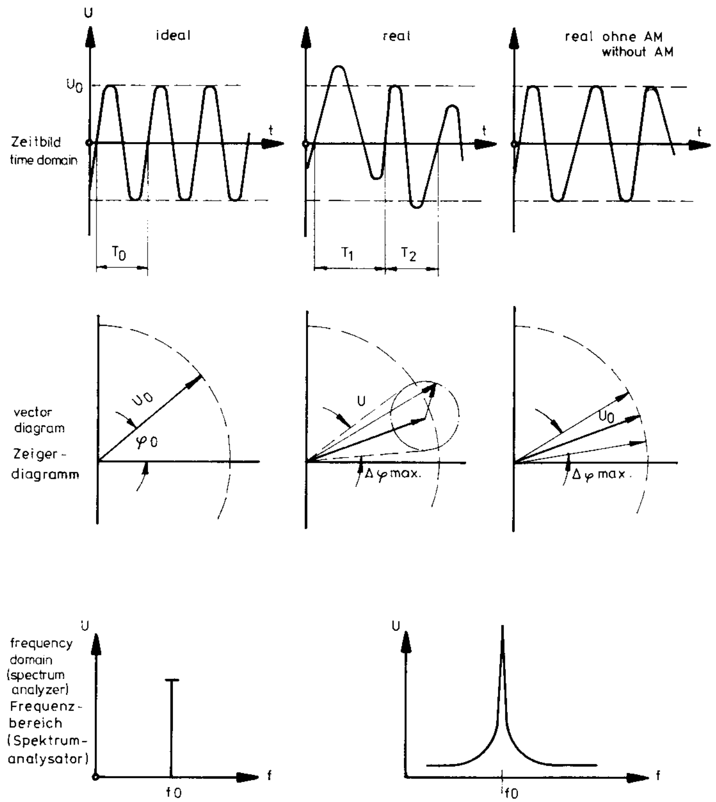


Fig. 2: Frequency stability in various presentations

Short-term stability is observed as »short-period« fluctuations of the selected frequency, or phase within a period of seconds. **Figure 2** shows this with the aid of various diagrams.

The upper diagram shows the oscillator signal in time domain, that is as can be observed on an oscilloscope. The center diagram shows the same effect in the form of a vector diagram: The amplitude is shown as the length of the arrow, and the phase angle by the angle φ_0. The sinewave oscillation corresponds to a rotation of the arrow, and the momentary value is the projection of the arrow onto the reference axis. The lower row shows the same effect in the frequency domain, that is when observing the signal on a spectrum analyzer.

An ideal oscillator possesses a purely sinusoidal output voltage of constant amplitude U_0, the point of the arrow will move at a constant speed in a circular direction, and the spectrum will consist of a single line with a frequency $\omega_0 = 2 \pi \times f_0$

$$U(t) = U_0 \times \sin(\omega_0 t + \varphi_0) \quad (1)$$

The output signal of a real oscillator – which has been rather exaggerated in Figure 2 – will exhibit short-term amplitude and phase (frequency) fluctuations that are more or less statistically distributed. In the case of the vector diagram, this means that the length of the arrow will fluctuate, and will also not be constant in its circular path, but will swing around the ideal position. In the spectral display, this will be observed as a wider line in conjunction with amplitude fluctuations.

Such a »noisy« oscillation can be determined mathematically as:

$$U(t) = [U_0 + \varepsilon(t)] \times \sin \underbrace{[\omega_0 t + \varphi(t)]}_{\Phi(t)} \quad (2)$$

This means that the amplitude U_0 is provided with a noise component due to a momentary value $\varepsilon(t)$, and the total phase $\Phi(t)$ by the component $\varphi(t)$.

Since an amplitude limiting takes place within the oscillator chain (oscillator, multiplier stages, receiver-IF in the FM-mode) it is possible to neglect the »amplitude noise« (right-hand column in Figure 2), and the output signal will be

$$U(t) = U_0 \times \sin[\omega_0 t + \varphi(t)] \quad (3)$$

The momentary frequency at any time-point t will amount to:

$$f(t) = \frac{1}{2\pi} \frac{d}{dt} \Phi(t) = f_0 + \frac{1}{2\pi} \frac{d\varphi(t)}{dt} \quad (4)$$

This equation shows the relationship between the frequency and the phase noise, which cannot be separated from another.

The output signal consists of a spectral line of finite width. Away from the main line, a virtually »white« wideband noise base exists, whereas the main line itself is increased in its direct vicinity by an increasing noise level. In addition to this, weaker lines may be observed from harmonics, and other spurious signals.

The deviations from the constant frequency can be split into two types: determinable ambient effects, and statistic effects. Both types cannot be clearly separated from another experimentally.

The ambient effects are influences of the operating voltage, the ambient temperature, mechanical and electrical loading, as well as aging of the crystals and other electrical components.

The statistic effects on the oscillator frequency are added to the ambient frequency variations. Such effects are caused by noise of the active circuit components, by additional noise of the components, including the crystal. The crystal noise originates from effects in the resonator structure, as well as from physical surface effects between crystal disk and electrodes. This error causes a dependence of the resonant frequency on the loading of the crystal, as well as variations of the electrical equivalent values on changing the exciting amplitude.

If one observes the common form of an oscillator, the main points where the noise is generated can be localized according to **Figure 4**:

The main components of an oscillator are:

- The amplifier for commencing and maintaining oscillation
- A limiter circuit, which is usually realized by the amplifier itself. (The task is to stabilize the maximum amplitude after commencing oscillation).

Furthermore:

- The resonant circuit – in our case the crystal, which may be provided with phase-shift, LC-reactive components.
- Finally, a buffer for isolation of the oscillator from the subsequent stages.

The level of the white wideband noise will be determined by the noise contribution of the buffer and subsequent amplifier stages; the increase of noise in the vicinity of the carrier frequency is formed by the selective oscillator network. The finite line width of the carrier itself is dependent on the Q of the crystal, the type of network, and the low-frequency noise characteristics (1/f-range) of the oscillator stages. The low-frequency noise can affect the oscillator in two different ways:

a) Directly on the frequency-determining parameters (crystal) which is called parametric noise

b) It can be multiplied and mixed up in the range of the carrier frequency indirectly via non-linearities.

The Q of the crystal is dampened in any oscillator circuit down to the so-called effective Q_{eff}, which means that the crystal and oscillator will possess a total Q_{eff} that may be (considerably) lower than the Q of the crystal on its own.

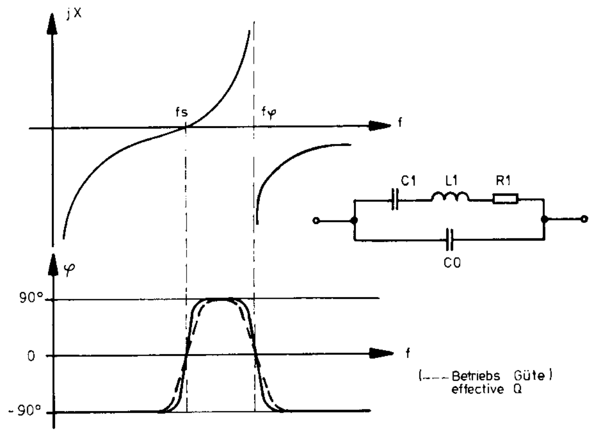


Fig. 3:
Reactive impedance and phase characteristic of quartz crystals

The relationship between phase noise and frequency noise is determined by the effective Q:

$$\varphi = \arc\tan\left(-2 \times Q_{eff} \frac{\Delta f}{f_0}\right) \quad (5)$$

The following results in the vicinity of the oscillator frequency f_0:

$$\frac{d\varphi}{df} = -2 \frac{Q_{eff}}{f_0} \quad (6)$$

Example: $Q_{eff} = 50000$, $f_0 = 96$ MHz

$$\frac{\Delta\varphi}{\Delta f} \approx -1 \times 10^{-3} \frac{rad}{Hz} \approx -0.05°/Hz$$

This means that the oscillator must offer constant phase conditions with an accuracy of 1/20° in order to obtain a short-term stability of 1 Hz (corresponding to 100 Hz at 10 GHz)!

Figure 3 illustrates equation (5); the dashed line is the phase characteristic of a dampened crystal.

What are the effects of short-term stability and phase noise in practice? Noise sidebands of a transmit signal will accommodate a frequency range in the vicinity of the transmit frequency whose width is dependent on the frequency multiplication factor, and they will also superimpose themselves on any weak signals that appear. In the case of receivers, these noise sidebands will be superimposed in the mixer on each receive signal and will cause a desensitization of the receiver in the vicinity of strong stations (1).

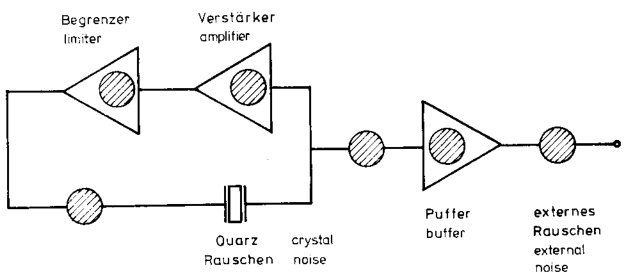


Fig. 4:
Noise sources in a crystal oscillator

2. MEASUREMENT OF THE SHORT-TERM STABILITY

2.1. Definitions, Measurement and Evaluation Magnitudes

2.1.1. Measurements in the Frequency Domain

If the noise spectrum in the vicinity of the carrier is studied more accurately, the following will be observed:

As was given in equation (4), frequency and phase variations are directly dependent on another. It is therefore sufficient when evaluating the short-term stability, to either measure the relative frequency variation (in ppm) with respect to a very stable reference frequency, or for the phase variations to be compared with a source having a quasi-stationary phase.

The mathematical evaluation is only to be explained briefly: The frequency deviations, (or phase deviations) are measured periodically in a certain measuring period, comparing only the variation from one measurement to the next. The statistic mean value of this product is called the auto-correlation function of the frequency or phase variation. From this one can obtain the spectral density of the relative frequency variation S_y (f), or the spectral densitiy of the phase noise $S\varphi$ (f) with the aid of an integral transformation.

The spectral density of the frequency stability and the phase variation are dependent on another according to the following equation:

$$S_y(f) = \left(\frac{f}{f_0}\right)^2 S_\varphi(f) \quad (7)$$

If one of these two experimentally measured spectral densities is drawn as a function of the frequency difference to the carrier, it is possible for the resulting noise spectrum to be differentiated according to type **(Table 1)**.

$$S_y(f) = \sum_{n=-2}^{2} a_n f^n$$

for $0 < f \leqq f_{max}$

$$S_\varphi(f) = v_0^2 \times \sum_{n=-2}^{2} a_n f^{n-2}$$

Figure 5 shows that one can draw one or more straight pieces between the measuring points when using a double-logarithmic display. These then have a slope of – 2, – 1, 0, etc. The physical nature of the individual components is summarized in Table 1. According to equation 7, the exponent of the frequency dependence is to the value of two less at S_φ (f) than at S_y (f).

Type of noise	S_y (f)	S_φ (f)
Statistic noise of the frequency	$a_{-2} \times f^{-2}$	$v_0^2 \times a_{-2} \times f^{-4}$
1/f noise of the frequency	$a_{-1} \times f^{-1}$	$v_0^2 \times a_{-1} \times f^{-3}$
White noise of the frequency Statistic noise of the phase	a_0	$v_0^2 \times a_0 \times f^{-2}$
1/f-noise of the phase	$a_1 \times f$	$v_0^2 \times a_1 \times f^{-1}$
White noise of the phase	$a_2 \times f^2$	$v_0^2 \times a_2$

Table 1

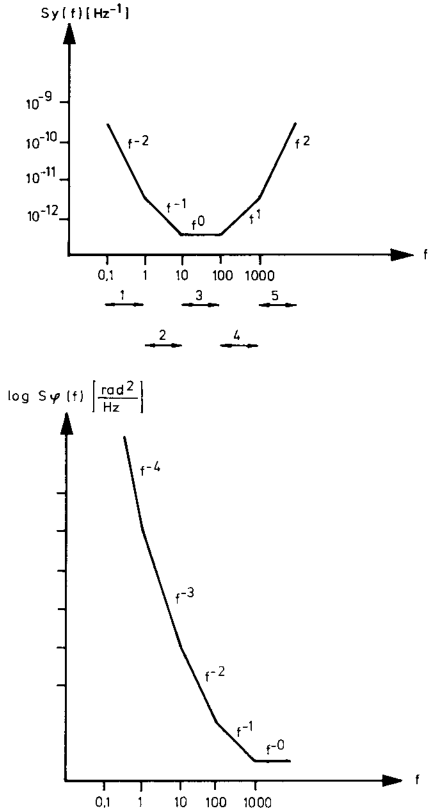


Fig. 5: Types of spectral distribution of frequency (S_y) and phase noise (S_φ)

Random noise of the frequency results in a noise spectrum proportional to f^{-2}. A noise behaviour f^{-1} represents the 1/f-noise of the frequency, and the white noise of the frequency will represent a constant contribution in magnitude S_y. It is equivalent to a random noise of the phase, i.e. $S_\varphi(f) \sim f^{-2}$.

The next line shows the 1/f noise of the phase and finally the white noise of the phase.

The most often used value for characterizing the phase noise (in the frequency range) experimentally, is the magnitude £ (f). This has the dimension Hz^{-1} and is defined as the ratio of the power at a certain point of a noise sideband to the total signal power, referred to a measuring bandwidth of 1 Hz (at the frequency $\nu_0 + f$, i.e. f Hz from the carrier).

$$£(f) = \frac{S(\nu_0 + f)}{P_{carrier} + 2 \times P_{sideband}}$$

Since the noise power is usually equally distributed to both sidebands, the following is valid for $\Delta_{\varphi noise} << 1$ rad:

$$£(f) = \frac{1}{2} S_\varphi(f)$$

Usually, the logarithmic value 10 log £ (f) (dB) is used instead of £ (f), and this is thus 3 dB lower than the logarithmic value 10 log S_φ (f).

If one measures in a bandwidth (analyzer bandwidth) of b Hz, the measured power will then be approximately 10 log b (dB) greater than at 1 Hz.

2.1.2. Measurements in the Time Domain

In the first part of this article we have described how the short-term stability can be characterized in a spectral display. The measurement of these magnitudes can be carried out with the aid of a spectrum analyzer.

It is, however, often easier to determine the stability with the aid of a large number of frequency measurements, or period duration measurements with the aid of a frequency counter. Measurements in the frequency and time domain are mathematically equivalent.

The most common value for characterizing in the time domain is the so-called Allan-variance. Using a frequency counter, one reads off a large number of frequency values f_i at predetermined intervals. The interval between two measurements is τ. For evaluation, however, only the relative individual differences

$$y_i = \frac{f_{i+1} - f_i}{f_i} \quad \text{(in ppm)}$$

between two subsequent values are formed. If one assumes that the gate time of the counter is very small with respect to the actual measuring period, the Allan-variance can be defined as follows with m measurements:

$$\sigma_y^2(\tau, m) = \frac{1}{2 \times (m-1)} \sum_{i=1}^{m^{-1}} (\bar{y}_{i+1} - \bar{y}_i)^2 \qquad (8)$$

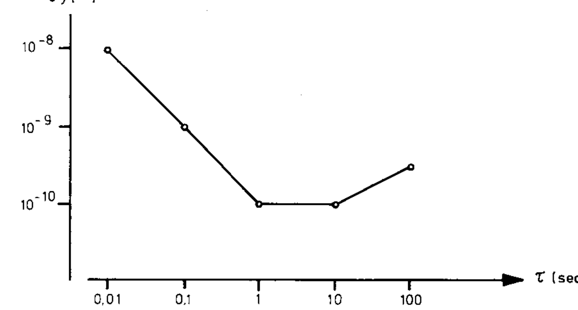


Fig. 6: Example of an Allan-variance measurement

The y-values are the relative frequency variations in ppm (10^{-6}).

This means that the frequency difference between measurement 2 and measurement 1 is formed, the difference (in ppm) is squared, and this value is stored. During the next cycle, the frequency difference between measurement 3 to 2 is formed and the difference square is summed up according to the previous result, which is followed by the third cycle, where the difference 4 to 3 is squared and summed up. After carrying out a sufficient number of measurements, the mean value of this sum of the difference squares is formed and divided by the duration of the measuring interval. This measurement is repeated for various measuring periods, e.g. 1/100 second, 1/10 second, 1 second, 10seconds, etc. and $\sigma_y(\tau)$ is drawn in double-logarithmic scale. This curve is a measure of the dependence of frequency on the measuring (integration) time from one measurement to the other.

In order to compare the characteristics of several oscillators to another, it is necessary for this type of measurement to be carried out at different measuring interval lengths.

Figure 6 shows the typical result of an Allan-variance measurement as a function of the counting time when drawn in a double-logarithmic scale.

2.2. Measuring Methods

A very good reference with respect to this section is given in (2): The NBS Technical Note 632, published by the National Bureau of Standards. This contains a detailed technical introduction with mathematical appendix, a detailed description of typical measuring systems for measuring the short-term stability with exact information for designing these, including circuit examples.

2.2.1. Frequency Domain (Spectrum Analysis)

The fundamental measuring principle is given in **Figure 7**.

A reference oscillator and the oscillator to be examined are converted to frequency zero in a balanced mixer. The conversion product is fed via a lowpass filter to a low-noise DC-amplifier, to which a phase-locked loop is connected. It is in turn loosely coupled to the reference oscillator so that it remains phase-locked to the oscillator to be examined. The phase-locked loop has a relatively large time constant, which means that only very slow frequency variations will be controlled. More rapid frequency variations will be seen as

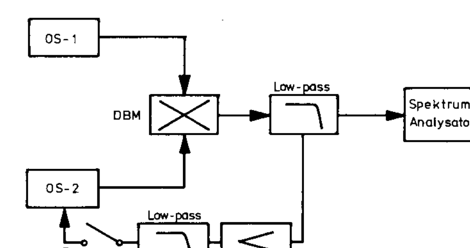


Fig. 7: Measuring principle sideband – phase noise

phase fluctuations, and will generate a noisy DC-signal at the output of the mixer. This can be taken from the output of the low-noise amplifier, and be examined using a spectrum analyzer suitable for low frequencies.

A measuring system was developed according to this measuring principle for examining the oscillators discussed in section 3; this system is constructed using conventional components. If only a coarse evaluation is required, it is sufficient for the noisy signal to be measured on an oscilloscope instead of using an expensive spectrum analyzer. The peak-to-peak amplitude of the noisy signal is a relative value, which can be used to carry out better/worse comparisons. The circuit diagram is given in **Figure 8**.

A peak rectifier or effective value meter can be connected in order to obtain a quantitative evaluation. It is possible to use a number of switchable, active bandpass filters and to measure their output signal instead of using the expensive spectrum analyzer. It is thus possible to evaluate the quantitative relationship between the various spectral components.

However, this method only allows two oscillators to be compared to another, whereas an absolute measurement on an oscillator is not possible. If oscillator 1 is considerably better (derived from a frequency standard), one will obtain a quasi-absolute measurement. If, on the other hand, virtually similar oscillators are compared to another, it will be necessary to deduct 3 dB from the measuring result, in order to obtain the phase noise of a **single** oscillator.

It is possible in this manner to construct various oscillator prototypes and to compare them to another.

The sensitivity of the system can be increased by carrying out the frequency conversion not at the frequency of the oscillator, but after multiplying the output frequency. A very elegant method is to use »successive 10/9 multiplication« (see **Figure 9**). This has the advantage that the frequency to be measured has the same frequency of the reference signal, whereas the noise sidebands are 10^n wider.

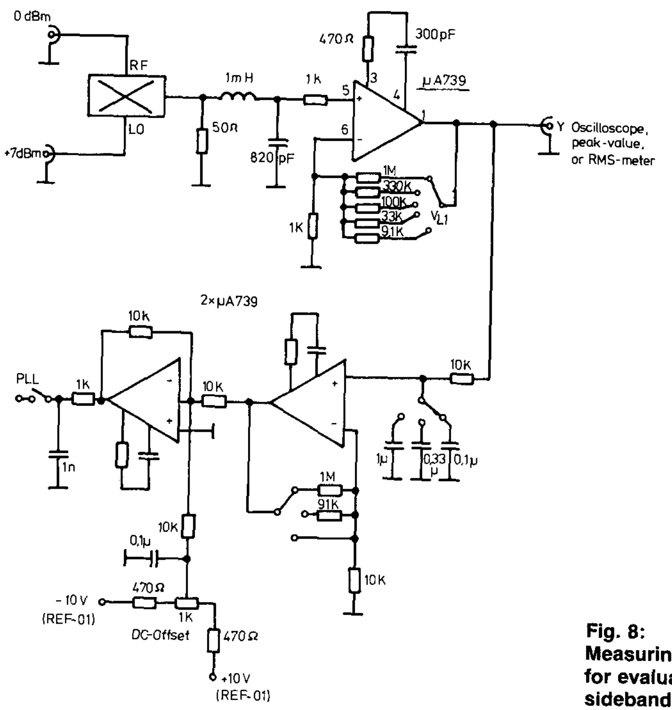


Fig. 8: Measuring circuit for evaluating sideband noise

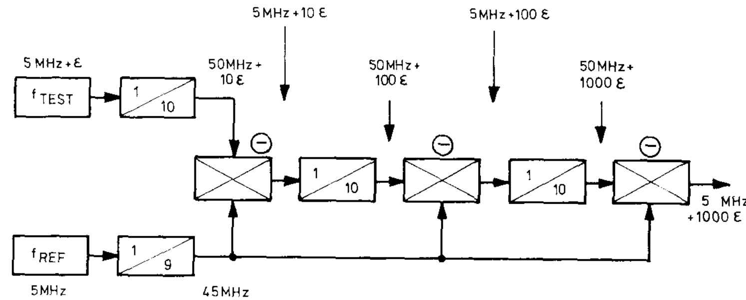


Fig. 9: Successive 10/9 multiplication principle

The quantitative measurement of the single-sideband noise £ (f) is made in the following manner:

a) The two oscillators are slightly detuned from another and the amplitude of the difference signal is measured, or adjusted to a reference value (0 dB). The peak-to-peak amplitude corresponds to the sensitivity of the measuring system in V/rad.

b) Both oscillators are locked-in opposite to another and the **effective** noise output voltage is measured at a spacing f from the carrier frequency. The voltage ratio is now converted to dB, and the bandwidth factor 10 log b is added to this. The result is the magnitude £ (f) in »dB/Hz«.

2.2.2. Measurements in Time Domain

Principally speaking, the time domain measurement observes the frequency variations of the oscillator directly using a frequency counter with the fastest possible count, and evaluating this statistically.

The time base of the frequency counter must, of course, have a considerably better short-term stability than the oscillator to be measured. Since a higher frequency resolution, – when using a directly measuring frequency counter – results in long measuring periods, it is usually necessary to make frequency measurements with a high resolution in short cycles, using a reciprocal counter and making period duration measurements.

a) Direct Measurement:
In the case of the direct measurement, the oscillator to be examined is directly connected to the frequency counter; the measuring results of the frequency counter are, for instance, printed out, or are directly evaluated using a fast processor. The most usual evaluation method is to use the Allan-variance.

b) Indirect Measurement:
In the case of the indirect measurement, the oscillator to be examined is mixed with a reference oscillator that is locked-in with the aid of a phase-locked loop (see frequency domain measurement); however, the control is so tight that the phase fluctuations are completely controlled. The correspondingly noisy output signal of the loop is amplified and fed to a voltage-frequency converter that converts the noisy DC-voltage into a rapidly-changing frequency fluctuation. This converted frequency can be measured with a frequency counter according to the method described in a), and evaluated.

One of the next editions of VHF COMMUNICATIONS will bring the final article.

An Extremely Low-Noise 96 MHz Crystal Oscillator for UHF/SHF Applications Final Part II

by B. Neubig, DK 1 AG

3. CONSIDERATIONS DURING THE DESIGN OF CRYSTAL OSCILLATORS WITH GOOD SHORT-TERM STABILITY

3.1. Selection of the most Favorable Output Frequency

The classical design of oscillators with a good short-term stability is based on a master oscillator at 5 or 10 MHz using precision crystals in their third or fifth overtone (6). This results in an extremely high frequency multiplication factor in the case of UHF and SHF frequencies. The noise component of the frequency multiplier stages adds a considerable amount to the total noise.

Due to technical advances in the production of crystals, it is now possible to use crystals having a higher resonance frequency. Fundamentally speaking, the Q of a crystal will increase on increasing the overtone (at the same frequency). It can be assumed that the product Q x f is a constant for a certain overtone. This means that the Q will deteriorate on increasing the frequency. The upper frequencylimitforcrystals onthe market is in the order of approximately 200 MHz. In professional technology, an output frequency range in the order of 100 MHz has been found to be most favorable. This means that our selection of 96 MHz represents the present state-of-the-art.

96 MHz crystals in fifth overtone are available readily in the larger HC-6/U or smaller HC-18/U case types. Comparison measurements madeby the author haveshown that the larger crystal has a more favorable Q. After carrying out a number of measurements, it was found that an HC-6/U crystal exhibited a mean Q of approximately 80 000, and 94 000 for HC-18/U. For comparison, crystals were also measured at their seventh overtone. This resulted in a mean Q of 106 000 in the case of HC-18/U crystals, which means approx. 13% higher Q than the fifth overtone. Precision crystals in a glass case exhibit a mean Q of 116 000 at the seventh overtone (see Section 4 for further details regarding the selection of a suitable type of crystal).

3.2. Selection of the Active Components

In the case of bipolar transistors, the noise is mainly determined by the base-emitter path. The noise of PNP-transistors is less than that of corresponding NPN-transistors. MOSFETs generate a high noise level, whereas the 1/f-noise dominates at low frequencies, and the thermal noise of the drain-source path dominates at higher frequencies. Junction field effect transistors generate less noise than bipolar transistors and MOSFETs (7). When field effect transistors are used, high-current types are preferable due to the larger, linear range and the low source impedance in a gate circuit. As can be seen in the measuring results given in section 4, transistor type P 8000 (now P 8002) is better than the approximately equivalent transistor BF 246 C.

3.3. Selection of the Most Suitable Circuit

The active stage of a crystal oscillator has two functions: The first is to provide a gain reserve for commencing and maintaining feedback conditions for oscillation, and the second is the limiting ofthemaximumpossible amplitude by reducing the gain at high amplitudes (saturation). In the case of an oscillator with good short-term stability, it is extremely important that both functions are separated from another and especially that the crystal will not »see« another component whose operating point will change during the commencement of oscillation, or after oscillation has been commenced. Otherwise, the fluctuating impedance in time with the RF-signal will cause a multiplicative conversion of noise sidebands.

Furthermore, a greatest possible feedback should be provided for the active stages, because a linear operating range, and a low-noise operation isonly possible inthismanner.

A decisive criterion for the selection of the oscillator circuit is the operating Q (see measuring results in section 4). A circuit that is able to fulfill these conditions was already described by the author in (6). The following information is to discuss the experience and measuring results encountered whilst obtaining a favorable design.

Figure 10 shows the principle of the circuit. As can be seen, it is a DC-coupled cascode circuit of two high-current field effect transistors. The basic oscillator circuit is a Collpits oscillator. If the source impedance of transistor T 1 is bridged capacitively, one will obtain a free-running LC-oscillator, where the feedback is generated using a capacitive tap on the resonant circuit L 1, C 1, C 2. Transistor T 1 operates in a common-source, and T 2 in a common-gate circuit.

The feedback isdesigned so that theoscillator will not oscillate on its own after removing the bypass capacitor at source 1. The resonant circuit comprising L 1, C 1, and C 2 is tuned to 96 MHz. If the crystal is now added into the source line, this will reduce the feedback of the source impedance due to the lower crystal impedance at the series-resonance frequencies (fundamental wave, third, fifth overtone etc.). The drain circuit ensures that oscillation is made at the required overtone. For this reason, it must have a sufficiently high Q. The ratio C 1 / C 2 determines the degree of feedback.

Both transistors operate in stable class A with a drain current of 40 mA. The limiting is made using two anti-phase Schottky diodes that are connected to a tap on the output circuit. The static capacitance of the crystal is compens-

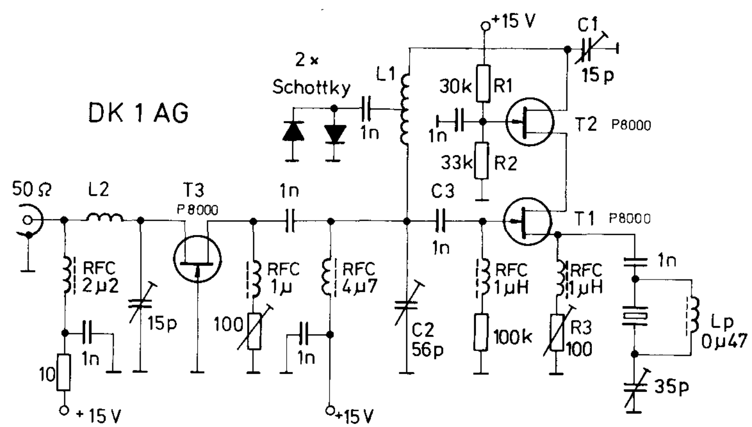


Fig. 10: Practical circuit of the 96 MHz oscillator
L 1: 6 turns of 1 mm dia. silver-plated copper wire wound on a 6 mm former, coil tap approx. at the center
L 2: as L 1, but without tap

ated for with the aid of L_p, in order to obtain a real crystal impedance, and a symmetrical phase run in the range of the series-resonance frequency of the crystal.

The cascode circuit equipped with the field effect transistors has characteristics similar to that of a tube pentode. The high internal impedance is characteristic, which means that the gain corresponds to the well-known tube formula

$$G = S \times (R_{in} \parallel R_{out}) \approx S \times R_{out} \quad (7)$$

The output impedance that can be realized in practice, is less than the internal impedance, which means in principle that a less than optimum matching is present. The same is valid for the drain of the first transistor. Its high-impedance output impedance is terminated with the low-input impedance of the source of the second transistor. In the opposite manner, source 2 will »see« the very high dynamic impedance of transistor 1, which is connected as constant-current source. This causes a very large feedback. A further feedback is provided by the source resistor R 3 in conjunction with the choke, which is in parallel with the series-resonance impedance of the crystal for RF.

The operating current of the cascode is adjusted with the aid of the source resistor; the voltage at the interconnection point of drain 1 and source 2 is adjusted to approximately half the operating voltage with the aid of the voltage divider R 1, R 2.

The two limiting diodes are removed for alignment, and the crystal is short-circuited. The free-running frequency is adjusted with the aid of the collector circuit to approximately 96 MHz. The compensating inductance L_p is aligned together with the crystal (after being removed) to 96 MHz with the aid of a dipmeter. The oscillator should operate in a stable manner after connecting the crystal and operating voltage. The oscillator frequency should only shift slightly on detuning the drain circuit, however, this should not be detuned too greatly in order to ensure that oscillation does not cease, or to avoid transcient problems on switching on. After adding the anti-phase diodes, the RF-amplitude should be limited to approximately half the value of the self-limiting oscillator by selecting the correct tap.

A subsequent buffer amplifier in a common-gate circuit (T 3) increases the output level to 18 dBm.

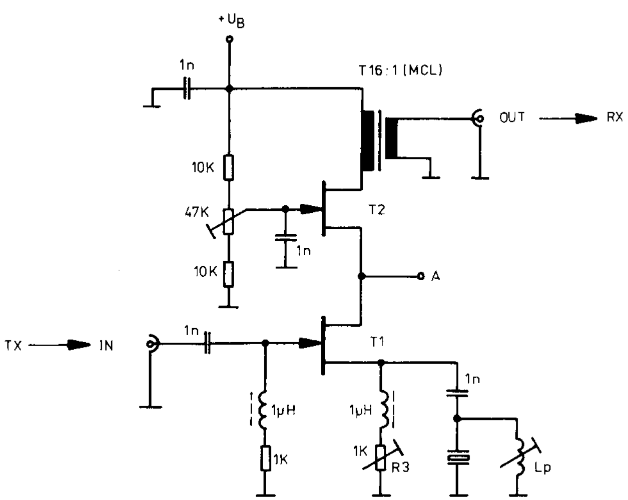


Fig. 11: Measuring circuit for determining the operating Q

4. MEASURING RESULTS

4.1. Operating Q as a Function of the Operating Point

The circuit was modified as shown in **Fig. 11** for measurement of the operating Q. The feedback line was cut, and a wideband transformer T 16-1 was used as output circuit. The 50 Ω signal from the signal generator of the spectrum analyzer was fed to gate 1, and the output signal fed to the 50 Ω input connector of the spectrum analyzer. If the crystal is bridged, the wideband gain can be read off from the spectrum analyzer.

On inserting the crystal, this gain will be fed back outside the resonances of the crystal, and one will obtain maximum gain at the series-resonance frequencies of the crystal, as well as at its spurious resonances. Inductance L_p can now be adjusted so that the resonance curve at 96 MHz is symmetrical.

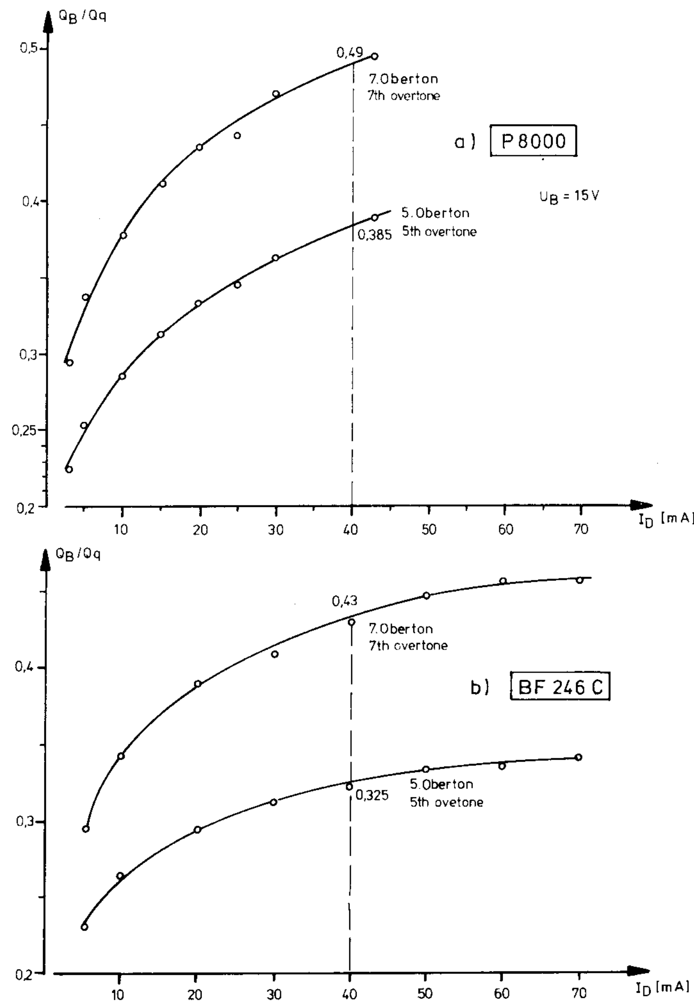


Fig. 12: Operating Q to crystal Q as a function of the drain current I_D

The voltage gain from gate 1 to drain 2 is now measured andthe bandwidth ofthe resonance curve is determined. This bandwidth is a measure of the operating Q.

$$Q_{op} = \frac{f_0}{\Delta f_{3dB}} \quad (8)$$

The operating Q as a function of the drain current was determined with a crystal operating in fifth overtone (Q = 89 300) and a crystal in seventh overtone in the glass case (Q = 127700) in conjunction with a transistor pair 2 x P 8000, and 2 x BF 246 C. The results are given in **Figures 12 and 13.**

Figure 12 shows the relationship between operating Q and crystal Q as a function of the drain current. The upper diagram is valid for transistor type P 8000, and the lower for transistor BF 246 C, and they show the results for both crystals. When using a BF 246 C, and an average fifth overtone crystal, it is possible to obtain a relationship of operating Q to crystal Q of between 23% and 33% according to the drain current. Using a transistor P 8000, one will obtain a value of Q_{op}: $Q_{cr} \approx 37\%$ when using the same crystal, and a drain current of 40 mA. The operating Q will be between 30 and 46% when using a seventh overtone crystal and a transistor BF 246 C, whereas a P 8000 will already reach the 50% mark at approximately 45 mA.

The operating Q is summarized as an absolute value as a function of I_D in Figure 13. One will notice the saturation effects in excess of 40 to 50 mA. For this reason, a drain current of 40 mA was used as operating point for the subsequent measurements. With this and with the inferior transistor and a fifth overtone crystal, an operating Q of approximately 28 000 was obtained, whereas approximately 62 000 (or more than twice) was obtained using a P 8000 and a seventh overtone crystal. Since the phase noise is mainly determined by the operating Q (see equation 1), it is easy to see that an improvement of more than factor 2 can be achieved by selection of the most favorable operating point and type of crystal.

Crystal Q is mainly dampened in the oscillator by two factors. The lowest component is the series connection of source-choke and R 3 that are parallel to the crystal, whereas the main component is the source-input impedance of transistor T 1. This is in series with the equivalent resistance of the crystal.

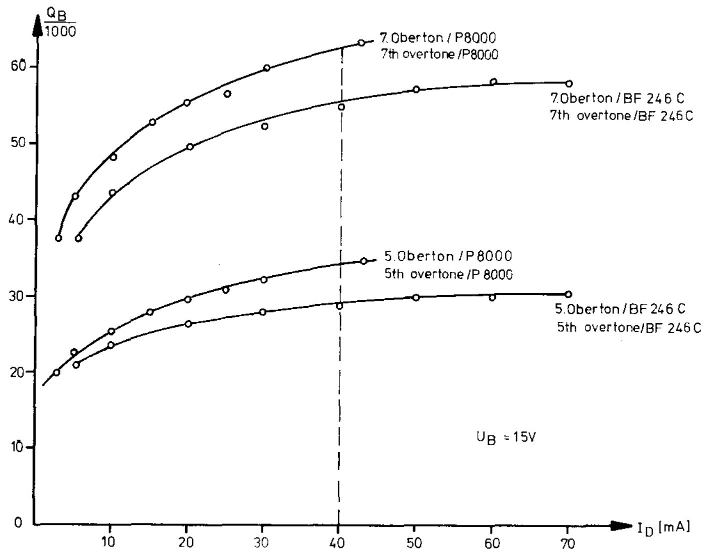


Fig. 13 Operating Q as a function of I_D

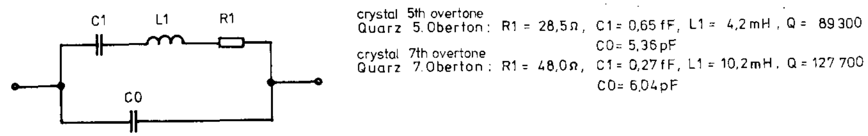


Fig. 14: Specifications of the measured crystals from Figures 12, 13, and 15

If one compares the equivalent data of both crystals (see **Figure 14**), one will see that the fifth overtone crystal possesses a lower Q. Furthermore, it exhibits a larger dynamic capacitance C 1 = 0.65 fF, which results in a series-resonance equivalent resistance of 28.5 Ω. The seventh overtone crystal exhibits a C 1 = 0.27 fF, which results in a resonance impedance of 48 Ω in conjunction with a Q of 127 700. The higher equivalent resistance of the seventh overtone crystal (in conjunction with a higher Q !) means that a certain source-impedance will deteriorate the overall Q less than when using a lower-impedance crystal. This means that the damping of the operating Q by the higher overtone crystal will be less than when using the fifth overtone.

Since the effective source impedance decreases on increasing the drain current, an increase of operating Q results on increasing the drain current. **Figure 15** shows this relationship with the aid of the effective source impedance determined with the aid of the above curves. This is in the order of 60 to 64 Ω for the BF 246 C at a drain current of 40 mA, and 45 to 50 Ω in the case of transistor P 8000.
Of course, a further improvement could be obtained when using a ninth overtone crystal, however, this would make the alignment of the oscillator more critical since the loop gain will be reduced considerably due to the higher source-feedback caused by the resonance impedance of the crystal.

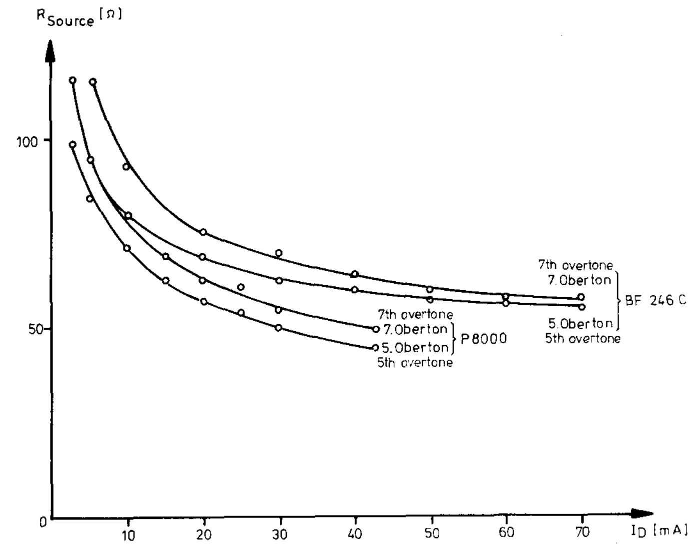


Fig. 15: Effective source impedance as a function of I_D

4.2. Most Favorable Output Impedance of the Oscillator Stage

A circuit as shown in Figure 10 was modified by removing the buffer stage, as well as interrupting the feedback by disconnecting C 3. The seventh overtone crystal is present in the source circuit. A low-level signal is injected to gate 1, and tapped off at drain 2 at high impedance (low-capacitance probe). The output level is measured at various terminating impedances.

Figure 16 shows the results:
The power gain (at $I_D = 40$ mA) reaches a saturation value of approximately 20 dB at $Z_T >$ 1 kΩ. The output power increases linearly up to $Z_T = 800\ \Omega$ (dynamic slope = 14 mA/V) and falls off above this.

In order not to absorb too much power due to circuit losses, it is advisable to set the impedance to ≈ 1 kΩ. The output circuit comprising L 1, C 1, and C 2 is also used as π-link in order to transform Z_T to the 50 Ω input impedance of the buffer. The feedback ratio 1:20 ≙ 13 dB provides a sufficient gain reserve (g = 20 dB) to ensure a stable commencement of oscillation.

4.3. Measurement of the Noise Sidebands

Two oscillators were built up according to the circuit given in Figure 10, and one of these was aligned to a fixed frequency of approximately 96 MHz with the aid of a trimmer. In the case of the second oscillator, the trimmer in the vicinity of the crystal was replaced by a pair of varactor diodes (see **Figure 17**), and locked to the frequency of the first oscillator according to the measuring principle shown in Figure 7.

The measurement was made using a professional laboratory system (Wandel & Goltermann FSM-1095) in both the frequency and time domaines. It is possible in this manner to measure up to 15 Hz from the carrier. The result is given in **Figure 18** as the noise spacing 10 log £ (f) referred to the carrier, and recalculated for a measuring bandwidth of 1 Hz (»dB_c/Hz«) for a sideband. The following noise sideband values will be seen:

Spacing = 100 Hz:	113 dB_c
Spacing = 1 kHz:	140 dB_c
Spacing = 10 kHz:	147 dB_c

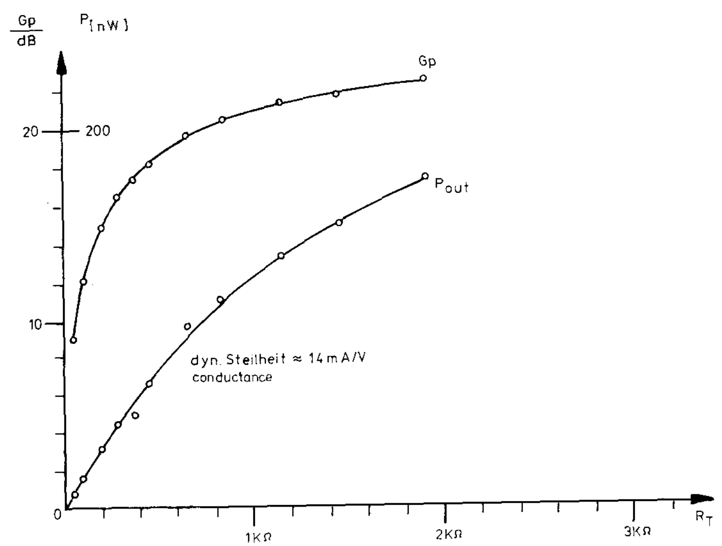


Fig. 16:
Power gain (dB) and output voltage as a function of drain-terminating impedance Z_T of the cascode stage (using seventh overtone crystal in the source circuit)

The curve can be subdivided in straight lines of different slopes corresponding to the behaviour of flicker noise, white frequency noise, and white phase noise, etc.

The importance of these very good values is to be discussed with the aid of an example:

After a frequency multiplication of 36,required for the 3456 MHz band, a value of − 113 dB_c means that an adjacent carrier spaced 3.6 kHz from the required frequency, will be superimposed with noise from the oscillator to the value of − 113 dB_c. If the carrier is S 9, corresponding to 50 µV or 34 $dB_{\mu V}$, the noise sideband will produce − 79 $dB_{\mu V}$ per Hz of bandwidth. At a bandwidth of 2.4 kHz (SSB), this corresponds to 10 log 2400 ≙ 33.8 dB more. This is equivalent to a noise floor of − 45.2 $dB_{\mu V}$ ≙ 5.9 nV, or 31 dB less than S 1. Of course, this is only valid when the unwanted carrier does not have any noise sidebands of its own.

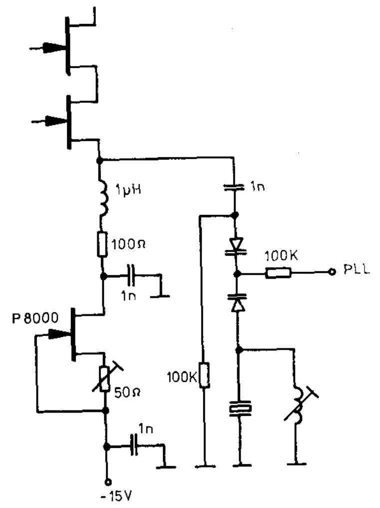


Fig. 17:
Improvement of the source circuit, and modification for pulling, or frequency modulation

5. PRACTICAL OPERATION AND IMPROVEMENTS

The operating current of 40 mA produces a considerable dissipation power, which is the price to be paid for obtaining a clean signal. The heat dissipation should be radiated effectively by a suitable mounting of both oscillator FETs. Furthermore, attention should be paid that it is not allowed to reach the crystal. It is therefore advisable for the crystal to be mounted outside of the oscillator case in order to ensure that no heating (frequency shift) occurs. For higher demands on the frequency stability, the crystal should be placed in a crystal oven, since even a crystal with a most favorable temperature response will drift by 1 to 2 ppm, equivalent to 100 to 200 Hz at 96 MHz for every 10°C.

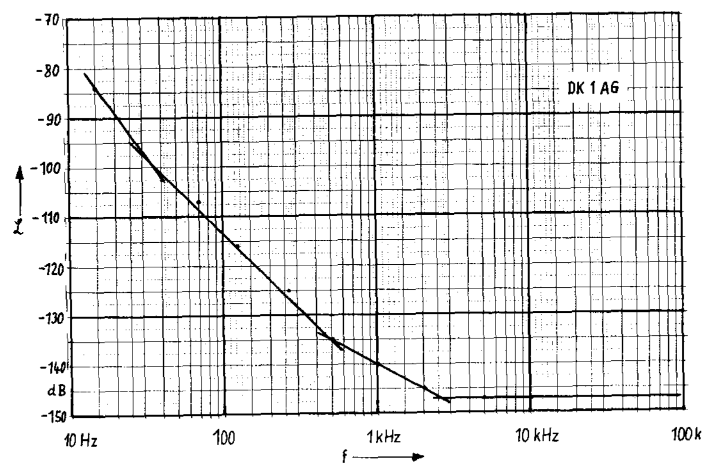


Fig. 18:
Noise rejection as a function of the frequency spacing from the carrier

If 40 mA is too high, it is possible to operate the circuit with 20 mA if a slight deterioration in stability is acceptable (see Figure 13).

The limiting diodes in the original circuit are Schottky types; if necessary. they can be replaced by noisier silicon diodes (1N 4151 or similar) or germanium diodes (AA 113 or similar).

The operating current of 40 mA is in the vicinity of I_D (0) in the case of some P 8000/P 8002 transistors, which means that the source leakage impedance, and thus the DC-feedback will be very low. This can be improved by providing a constant current source as shown in Figure 17, which can be fed from − 15 V. It may be sufficient when the source resistor of 390 Ω / 1 W is connected to − 15 V. Figure 17 also shows how a pair of varactor diodes can be inserted to allow frequency modulation. The pulling range that can be obtained is given in **Figure 19**. A linear range of ± 5 ppm (≙ ± 500 Hz) can be obtained with a pulling voltage of between 1 V and 7.5 V in conjunction with the seventh overtone crystal; approximately 2.5 times this is obtained in conjunction with a fifth overtone crystal.

The bias voltage for these varactor diodes must be taken from a low-noise voltage source, such as from the 10 V stabilizer REF-01, manufactured by Precision Monolithics (PMI) or a suitable, filtered battery.

A further increase of the phase purity can be achieved using a simple crystal filter as shown in **Figure 20**, which is connected to the output of the buffer. A cheap fifth overtone crystal is suitable, and one will obtain a 3 dB bandwidth of ± 3 kHz. When using a seventh overtone crystal, one will obtain a lower bandwidth of ± 1.4 kHz.

The alignment is made very simply by adjustment of C 2 and C 4 for maximum output level.

The obtained selectivity curve is given in **Figure 21**. The filter has no effect on the directly adjacent channel, but will attenuate the wideband noise spectrum by approximately 20 dB if constructed carefully.

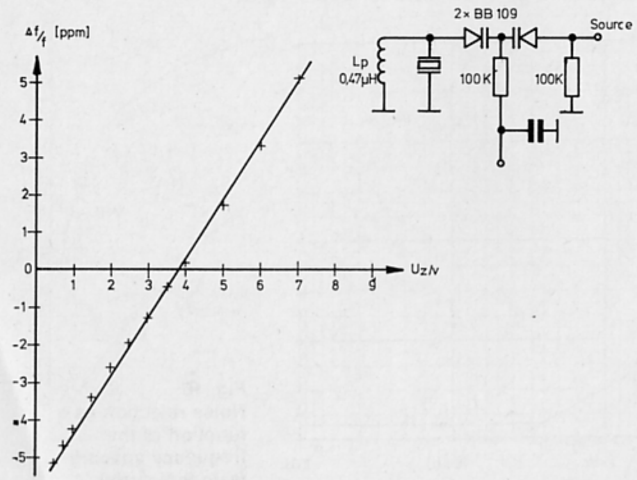


Fig. 19:
Frequency pulling characteristics of a seventh overtone crystal

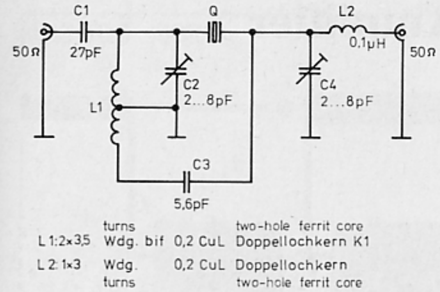


Fig. 20:
Crystal filter for improving the phase noise

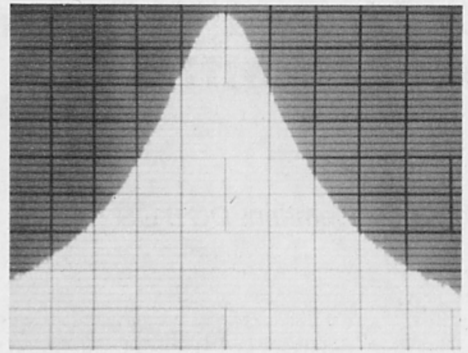


Fig. 21:
Selectivity curve of the crystal filter shown in Fig. 19 using a seventh overtone crystal
X = 1 kHz/div.; Y = linear

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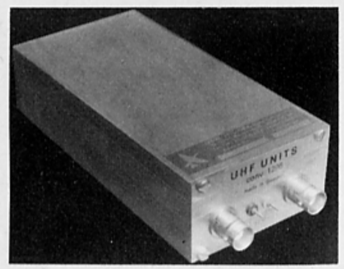

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