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Especially Covering VHF, UHF and Microwaves

VHF COMMUNICATIONS

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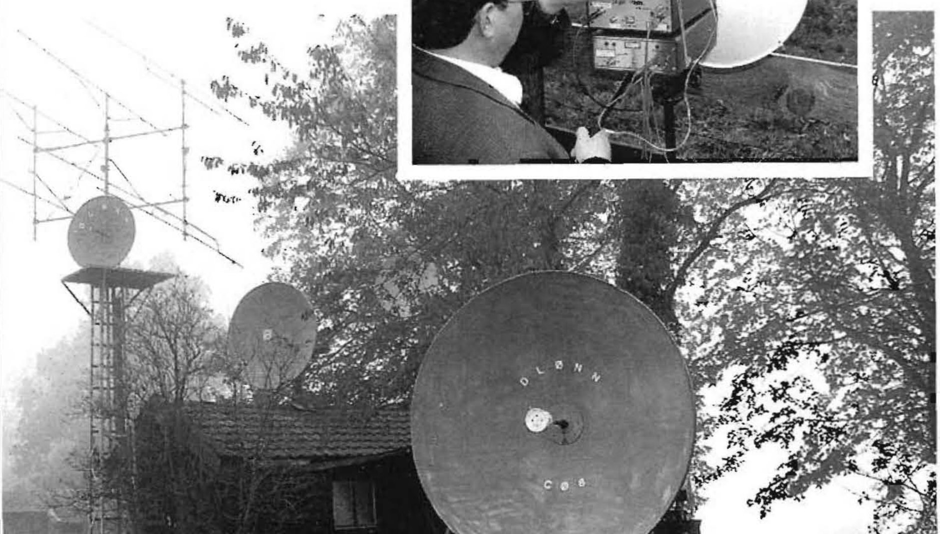

The DLONN Club Station at Holzkirchen

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Dipl.-Ing. J.v.Parpart

Carrier Suppression in the Ring Mixer

Electronic news technology can not be imagined without the mixing or modulation process. The so-called ring mixer was already in use in the 'thirties in the telephone technology sector. And even today, the ring mixer is still an up-to-date component, distinguished by high modulability and good reproducibility of technical data.

Depending on the application, it can cause interference if the oscillator signal cross-talks at the mixer output (inadequate carrier suppression). This problem is comprehensively discussed below and a solution is put forward.

1. RING MIXER THEORY

A ring mixer has three gates. At one gate, an oscillator feeds in the frequency: fo. At another gate, an input signal with the frequency fe is provided for the mixer. The mixer continuously switches the polarity of the input signal, using the polarity switching frequency, fo, and a keying ratio of 1:1. The polarity-switched signal appears as an output signal at the third gate of the mixer. It can be calculated what spectrum fractions are involved (Fig.1).

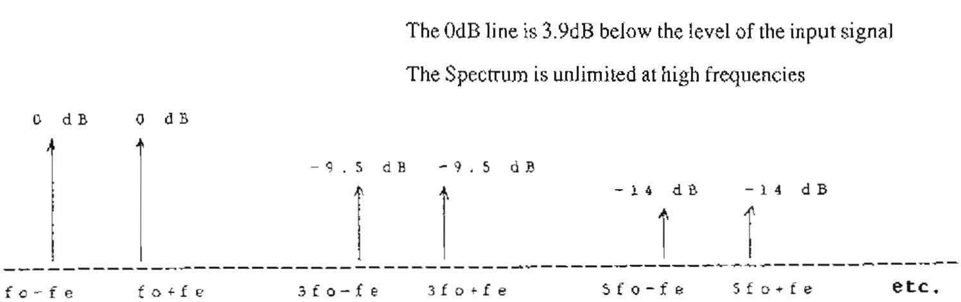


Fig.1: Spectrum fraction of a Mixer signal

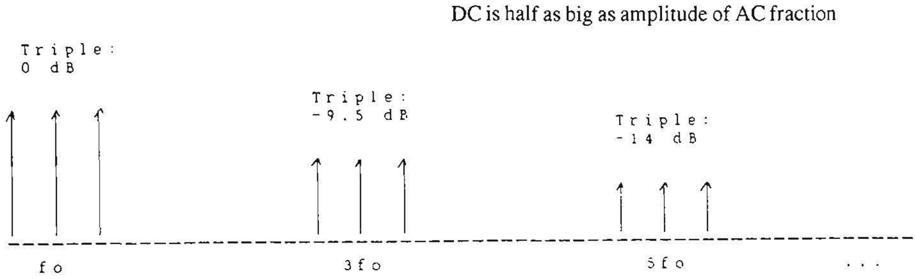


Fig.2: DC as Modulation signal

Among other effects, the ring mixer changes the frequency of the input signal. Fe becomes fo + fe or fo - fe. The conversion loss amounts to 3.9dB (that's physics, it can never get any better).

Two of the three gates of the mixer are coupled to a transformer. DC fractions can thus not be transmitted. One of the gates occupies a special position. Its transmission range begins at 0 Hz (DC).

Let us assume that the input signal is fed in at the DC-coupled gate. The spectrum at the mixer output changes if the input signal has a DC fraction. Carrier oscillation comes into play (frequency: fo), together with its odd multiples. If the DC fraction of the input signal is half as big as the amplitude of the AC fraction, then the carrier oscillation and the side bands are the same size (Fig.2).

The case outlined corresponds to a degree of modulation of 200%. Doubling the DC raises the carrier by 6dB. Halving it lowers the carrier by 6dB.

The carrier oscillation and the oscillator signal are in phase or in counter-phase to one another, depending on the DC polarity.

2. CARRIER SUPPRESSION IN PRACTICE

If, for example, we develop an SSB transceiver, then the low-frequency signal will be mixed with a 9 MHz oscillator signal. The carrier is unwanted. So we take a ring mixer and keep it DC-free.

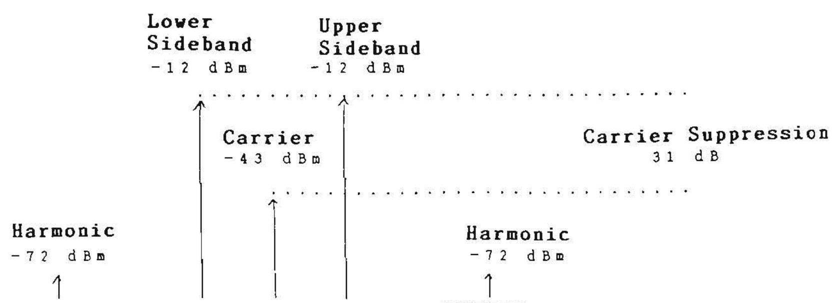


Fig.3: Damping value in Ring Mixer

Theoretically, there are no limits on the degree of carrier suppression, but how do things look in practice?

A standard ring mixer, for example, requires an oscillator level of 7dBm. The mixer manufacturer specifies, for example, that the oscillator signal has to be suppressed at 50dB, i.e. there are thus -43dBm at the mixer output.

The wanted signal - i.e. the converted low-frequency signal - at the mixer output is not particularly large. Otherwise non-linear distortions arise. And why should we make problems for ourselves as early as the mixing stage? So we select the wanted signal to be about - 12dBm. The harmonic distortion attenuation (k3) is then about 60dB (Fig.3). Narrow limits are set to any increase in the wanted signal, for it is known that the harmonic distortion attenuation (k3) is impaired by 2dB if the level is raised by 1dB.

What degree of carrier suppression takes place? Well, only 31dB are left over. That is the difference between the level of the wanted signal and the carrier.

3. PERFECT CARRIER SUPPRESSION

What is to be done to make the unwanted carrier disappear from the spectrum? One suggestion is to add a signal of the same size but in phase opposition. The sum then immediately becomes zero. The goal is achieved.

We know nothing about the couplings and asymmetry effects which (in contrast to pure theory) have caused the carrier to emerge. We can only start from the assumption that the phase length and the amplitude of the carrier are haphazard but stable.

According to the laws of vector calculation, the carrier oscillation can be broken down into two components. One component is aligned in the same direction as the oscillator signal (0° or 180°). The other component is orthogonal to the oscillator signal (90° or -90°). Both components must disappear for perfect carrier suppression.

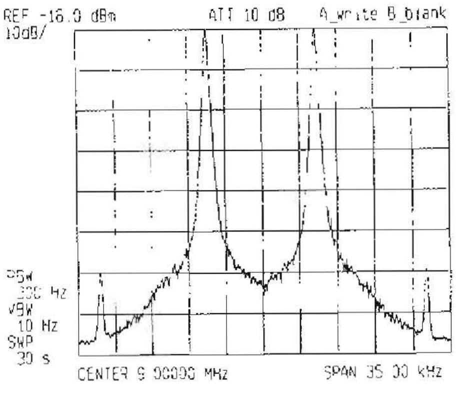


Fig.4a: Modulated signal without Carrier, fo

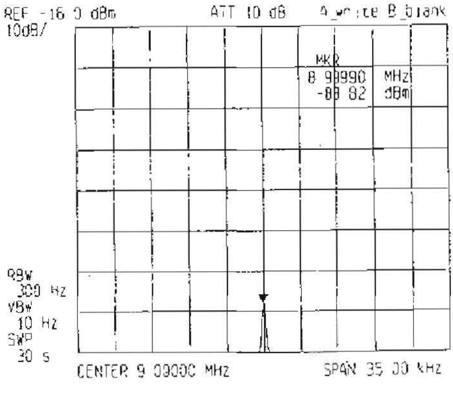


Fig.4b: Carrier only becomes visible without modulation

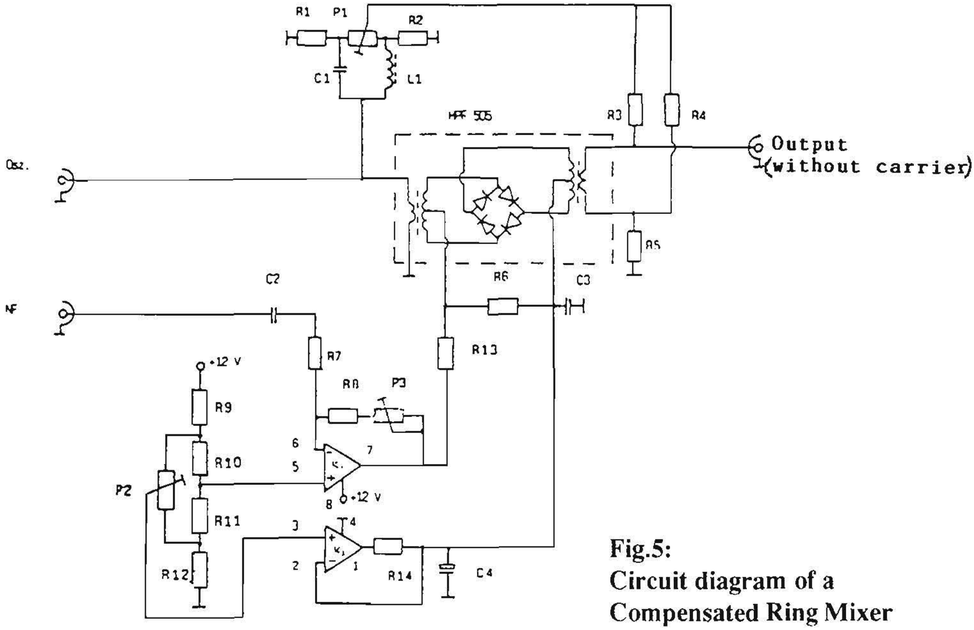


Fig.5:
Circuit diagram of a Compensated Ring Mixer

The component which is in phase with or in phase opposition to the oscillator signal can simply be compensated for by means of a DC voltage at the DC input of the mixer. The polarity and level of the DC must be adjustable.

The orthogonal component can be compensated for if a signal displaced 90° is obtained from the oscillator signal and added to the mixer output signal. The phase length (90° or -90°) and the amplitude of the signal must be adjustable.

If the process is carried out skilfully, two potentiometers are sufficient for tuning.

4. CIRCUIT FOR A 9 MHz MIXING STAGE

The proposed circuit is intended to mix a low-frequency signal with a 9 MHz oscillator signal. The crucial element is a ring mixer (HPF 505), which is linked into the compensation circuit referred to above. The level for full modulation has been selected in such a way that a harmonic distortion attenuation (k3) of 60dB is set. The harmonic interval of the low-frequency generator must be extremely good for this measurement. The UPA3 (R & S) was used. The following test certificates show that there is scarcely anything to be seen of the carrier. It can not be discovered until the low-frequency generator is switched off (Figs.4a, b).

And now for the circuit.

4.1. Power supply

The circuit requires 12 Volts, 20mA. A standard fixed voltage controller is used.

4.2. Low-frequency signal path

The low-frequency signal should be at a level of 6dBu (4.4 Vss). Deviations of +/- 3dB are permissible. The potentiometer, P3 (Fig.5), is set in such a way that there are 8 Vss at pin-7 of the operational amplifier (IC1). The lower frequency limit is determined through R7 and C2 and comes to about 30 Hz.

The operating point of the operational amplifier IC1 (1/2NE 5532) is mainly determined through the voltage dividers, R9 and R12, and lies in the middle of the useful range of modulation.

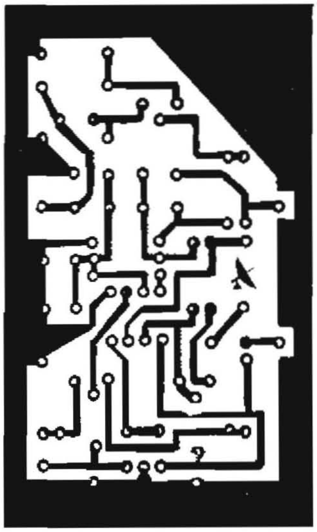


Fig.6: Layout

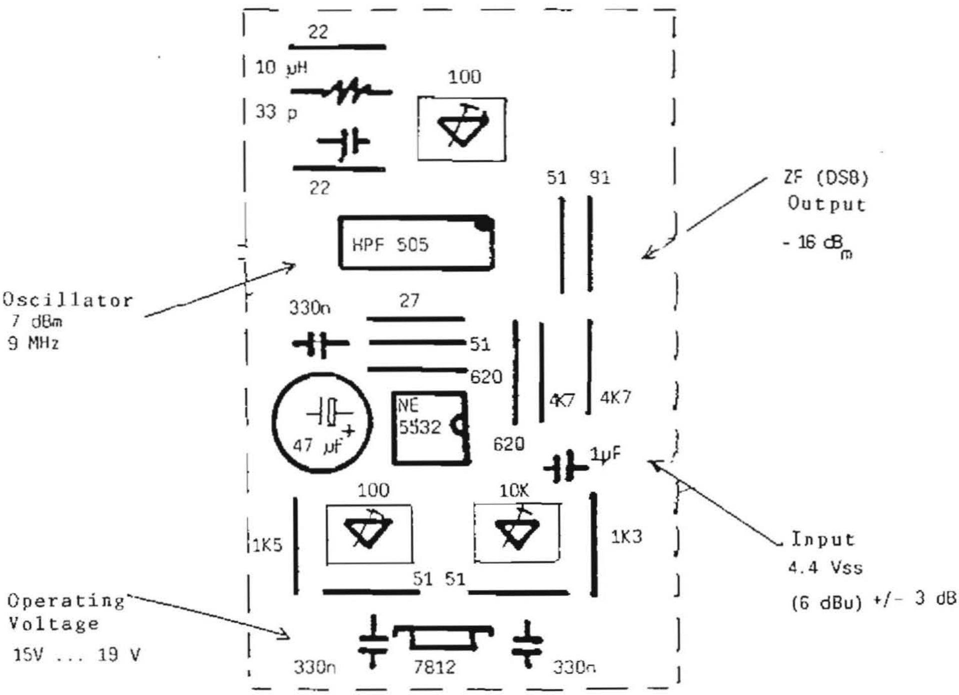


Fig.7: Assembly plan for the Compensated Ring Mixer

The operating point of the operational amplifier IC2 (1/2NE 5532) is set using the potentiometer, P2. Should the loop of the potentiometer be in the central position, then the operational points are the same for both operational amplifiers. With a right-hand or left-hand deflection, there are differences amounting to 100 or -100 mV.

The consequence is that between pin-7 of IC1 and the capacitor, C4, there is a signal with an AC fraction and an adjustable DC fraction. This signal is fed into the DC input of the mixer through R13. The resistance, R6, provides for a termination of this input corresponding to the surge impedance.

4.3. The path of the oscillator signal

The oscillator signal (9 MHz, 7dBm) is connected to the mixer input. A small part of the signal is decoupled through C1 and L1. The reactive impedances are equal and amount to approximately 500Ω. At R1 and R2 are voltages which are phase-displaced by (almost) 90° or -90° in relation to the oscillator signal.

If the loop of the potentiometer (P) is in the central position, then (almost) no signal is picked up. If the loop is rotated out of centre, a signal can now be picked up - always depending to which side we rotate - which is phase-displaced by 90° or -90° to the oscillator signal. The amplitude can be as high as 2% of the oscillator signal (right-hand or left-hand deflection).

The compensation signal must now be inter-connected with the signal at the mixer output again. In order to obtain a high degree of decoupling, a bridge circuit is used. If the intermediate-frequency output is terminated at 50Ω (ZW), then the mixer output is in the zero branch of a balanced bridge: R4 : R5 is as R3 to ZW. The consequence is that the compensation signal does not get to the mixer, but only to the intermediate-frequency output.

4.4. The path of the intermediate-frequency signal

If the intermediate-frequency output is terminated at 50Ω (ZW), then the potentiometer, P1, is located in the zero branch of a balanced bridge. R5 : ZW is as R4 to R3. The consequence is that the signal at the mixer output does not reach the potentiometer, but only the intermediate-frequency output. Irrespective of the potentiometer setting, the mixer output is always matched and the signal at the intermediate-frequency output remains constant (both side-bands: -16dBm).

The bridge circuit damps the mixer output signal by 3.8dB with the given dimensions.

4.5. Tuning

The low-frequency signal at pin 7 of IC1 is set to 8 Vss using P3. At the intermediate-frequency output, both side-bands are then -16 dBm.

The 0° / 180° cross-talk is compensated for using P2.

The 90° / -90° cross-talk is compensated for by P1.

4.6. Parts list

IC1: NE 5532	Mixer: HPF 505
R1, R2: 22Ω	R3: 91Ω
R4, R6: 51Ω	R5: 27Ω
R7, R8: 4.7Ω	R10, R11: 51Ω
P9: 1.3 kΩ	R12: 1.5 kΩ

R13: 620Ω	R14: 620Ω
P1: 10 kΩ	P2: 100Ω
L1: 10µH	
C1: 33 pF	C2: 1µF (foil)
C3: 330 nF	C4: 47µF

5. GOOD AND BAD AMPLITUDE MODULATION

Let us assume that an amplitude modulator is to be developed. It is proposed to connect a signal with a DC fraction to the DC input of the mixer, as above.

With full modulation (100% AM), the carrier is 6dB larger than the side bands. The DC must thus be as large as the amplitude of the AC fraction.

One obvious question is this. Why should we be thinking about carrier suppression if we still want a large and powerful carrier to be present in the spectrum? The fiy in the ointment is that the oscillator signal cross-talks at the mixer output. The 0° or 180° component of the cross-talk does not cause any other interference. It merely changes the degree of modulation. It's the 90° or -90° component which is dangerous!

An oscilloscope would show what is happening over time. It is not possible to achieve 100% modulation. A residual carrier always remains (namely, the 90° / -90° component of the cross-talk). To put it another way, with a high degree of modulation the envelope curve is deformed. An envelope curve de-modulator could not win the signal back free of distortion.

A second effect now arises. The carrier is phase-modulated. Why this is so can easily be explained by means of a vector diagram. The amplitude-modulated signal (horizontal axis) displays instantaneous values which vary in size. The phase length of the resulting vector is altered by the 90° or -90° component of the cross-talk.

It is relatively simple to eliminate the error (Fig.9). It is necessary to compensate for the 90° / -90° component of the cross-talk. To this end, as already indicated, part of the oscillator signal is decoupled and added to the mixer output signal after a phase displacement of 90° or -90°. For tuning, the AM signal (e.g. degree of modulation 80%, low frequency 1 kHz) is fed into a commercial FM or PM demodulator and the percentage interference modulation is measured.

6. THE INTER-CARRIER PROCESS

The so-called inter-carrier process is known from television engineering. The fact is made use of that 5.5 MHz or 5.74 MHz below the FM-modulated tone carrier there is a crystal-stabilised carrier - namely the image carrier.

For the purposes of selection (and Nyquist filtration), the image and sound carriers in home receivers are switched over to the intermediate-frequency position (38.9 MHz, 33.4 MHz/33.16 MHz). The inter-carrier modulator (depending on the circuit design, it may be formally described as a "quasi-parallel sound de-

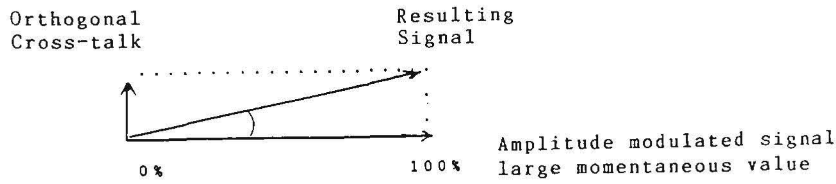


Fig.8: Vector diagram of an AM signal

modulator"), then forms the differential frequency, 5.5 MHz or 5.74 MHz, by mixing the sound carrier and the image carrier (in the intermediate-frequency position). FM demodulation then follows.

Oscillators were previously used for conversion to the intermediate frequency position. Today (cheaper) synthesisers are used. Cheap oscillators are known not to be stable over the short run, i.e. they have a percentage interference modulation. The image carrier and the sound carrier suffer the same percentage interference modulation through conversion. But the differential frequency (5.5 MHz or 5.74 MHz) does not alter. The advantage of the inter-carrier process lies in the fact that the percentage interference modulation of the converter oscillator has no disadvantageous effect on the quality of the sound transmission.

It is a very similar story if the home receiver is connected to the cable network. In certain circumstances, the converter also operates in the terminal with a poor

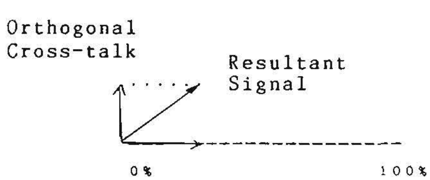


Fig.9: Amplitude Modulated signal small momentaneous value

synthesiser (present designs use a reference frequency of 781.25 Hz; this corresponds to a channel grid of 50 kHz with fixed pre-division by 64). The percentage interference modulation of such converters is considerable. De-modulation after the parallel sound process would lead to insufficient external voltage and noise voltage intervals. So the parallel sound receiver has finally died. Liberal regulations permitting CATV converters are to blame.

Back to the inter-carrier process. There are two important requirements which the image transmitter has to meet. The image carrier must always be available at sufficient amplitude (in this country: 10% residual carrier) and the image carrier may not be frequency-modulated or phase-modulated.

The inter-carrier modulator can not distinguish whether the phase/frequency of the image carrier or the sound carrier alters. Wherever the change comes from, it is demodulated.

The main reason for unsatisfactory quality in television sound is the so-called image modulator. It is just the same as with a "standard" amplitude modulator. The 90° or -90° component of the oscillator cross-talk brings about the phase modulation of the image carrier.

Really outstandingly good external and noise voltage values for television sound can be obtained only if the image modulator is tuned at minimum 90° or -90° oscillator cross-talk. A simple test is as follows: modulate with black image, intermediate-frequency output from image modulator to commercial intermediate-frequency de-modulator, and measure percentage interference modulation.

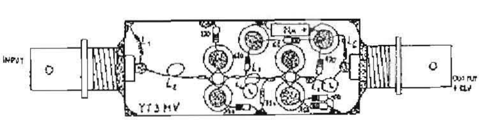
For the rest - you still keep finding the false opinion in the relevant literature that the residual side-band filter of the transmitter is responsible for the phase modulation of the image carrier. It is naturally indisputable that every one-sided band transmission involves a phase modulation. But the residual side-band filter does not cut in until approximately 750 kHz with video frequencies, and this is outside the audible range.

7. CONCLUSION

It is possible to perfect the carrier suppression in a simple manner using ring mixers. The process proposed is based on the fact that both components of the cross-talk are to be compensated for: the component which is in phase with or in phase opposition to the oscillator signal and the component which is orthogonal to the oscillator signal. The results obtained were discussed with reference to a 9 MHz mixer step.

There are circuits which can be decisively improved if the orthogonal components of the cross-talk are compensated for. These include the amplitude modulator and the image modulator used in television engineering.

A patent was applied for the operating principle of the compensation circuits in 1987 by the then employers of the author (P 37 32 171). For commercial use, then, a check should be made as to whether corresponding patent rights exist.



Dr. Ing. Jochen Jirmann, DB1NV

Theory and Practise of the Frequency Synthesiser Part-2

The basic procedures for frequency synthesis were put forward in the first part of this article. Now Part-2 deals with the elements of a synthesiser, namely the phase discriminator, the variable-gain amplifier, the voltage-controlled oscillator and the reference frequency generation equipment; filters and mixers should certainly need no further explanation and are thus excluded.

5. PHASE AND FREQUENCY DISCRIMINATION

The most important element of a phase control loop is the phase discriminator; it generates an output voltage which, in the ideal case, is proportional to the two signals fed in. A phase discriminator is nothing other than a mixer, which mixes down to an intermediate frequency of "zero", and thus supplies an oscillating DC voltage at its output depending on the phase difference.

So any component with a non-linear characteristic can be used as phase discriminator. In order to find a phase discriminator which can be used in practise, the unit must be optimised to specific properties, which has resulted in the following basic types:

5.1. The analogue multiplier

An analogue multiplier is nothing other than a double balance mixer, and forms the product from two applied signals. If two sinusoidal input signals are present with the frequencies f_1 and f_2, then the output voltage, as the product of the input voltages, can be expressed as follows:

$$U_A = U_1 \sin(2\pi f_1 t) \cdot U_2 \sin(2\pi f_2 t) \qquad (3)$$

When the basic equations for multiplying sin functions are used, we obtain:

$$U_A = U_1 \cdot U_2 \cdot \tfrac{1}{2}(\cos(2\pi t(f_1 - f_2)) - \cos(2\pi t(f_1 + f_2))) \qquad (4)$$

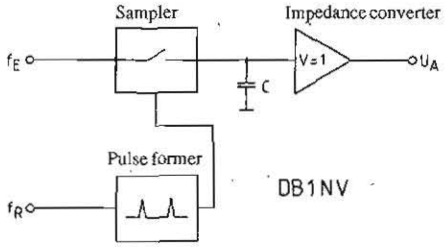


Fig.1: Principle of Sampling Phase Discriminator

This is the well-known "mixer equation" for an ideal mixer; only the total frequency and difference frequency of the input signals are obtained. Now, if the two signals have the same frequency and have a phase displacement of the output voltage of the multiplier has a DC fraction which depends on the phase position, and a fraction with a doubled frequency. A succeeding low-pass filter suppresses the fraction with the doubled frequency:

$$U_A = U_1 \cdot U_2 \cdot ½(\cos\varphi - \cos(4\pi f_1 t + \varphi)) \qquad (5)$$

As a phase comparator, the analogue multiplier has a number of advantages and disadvantages which are briefly listed here:

Since the multiplier is a linear component, it reacts insensitively to input signals which still have frequency components other than the wanted signals, unless the resulting mixed product also contains a DC fraction. It is thus possible to synchronise a PLL to a weak wanted signal in the presence of strong unwanted signals. This has the disadvantage that our output voltage depends not only on the phase difference but also on the amplitude of the input signals.

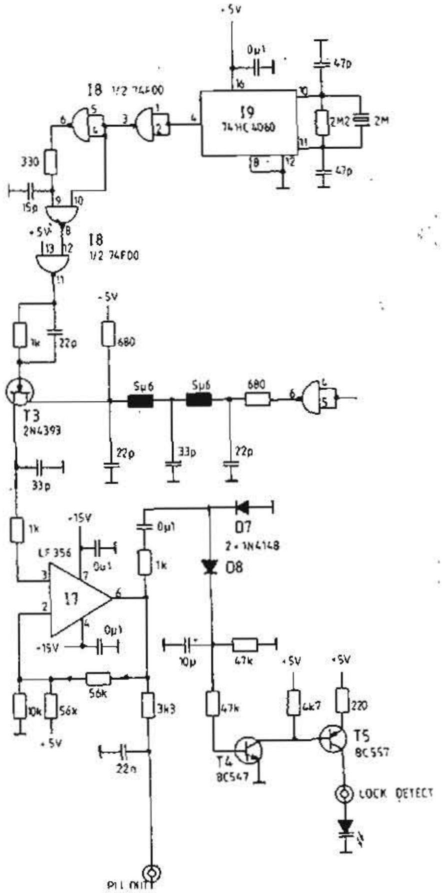


Fig.2: Section of PLL assembly DB1NV 008

The multiplication causes a strong unwanted signal to arise at the double frequency, which must be filtered out.

If the symmetry of the circuit is insufficient, the input signals can also cross-talk directly onto the output.

Passive phase discriminators in the form of diode ring mixers can be created for anything up to the highest frequencies.

The analogue multiplier operates satisfactorily only with input signals with a symmetrical pulse-width repetition rate.

5.2. The sampling phase discriminator

The sampling phase discriminator or scanning discriminator is a variant of the analogue multiplier; in mathematical terms, the input signal is multiplied by a pulse sequence with an amplitude of 1 and infinitely brief intervals (a so-called Dirac sequence). Fig.1 shows the layout in practise. One input signal is applied at a switch which is linked to a storage capacitor from which the output voltage is taken off at high resistance. The second input signal is converted into a sequence of pulses which controls the switch. The momentary value of the input voltage is scanned at each pulse (sampling) and intermediately stored in the capacitor. With input signals with the same frequencies, different DC values are obtained, depending on the position of the scanning point on the sine curve. If the frequencies are different, the differential frequency is obtained.

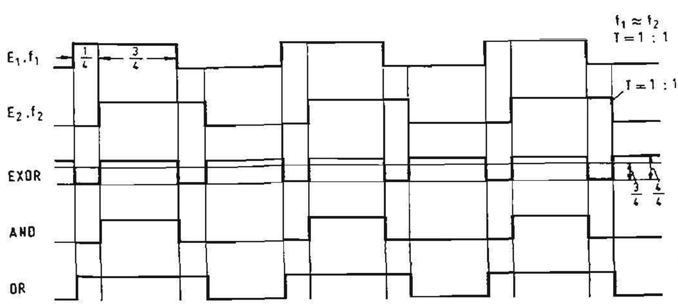
The characteristic properties of the scanning discriminator are:

At identical input frequencies, it supplies an almost pure DC, which contains only slight high-frequency spurious components.

The input signal which is applied at the switch must be sinusoid (sine or triangle), so that a continuous relationship exists between the phase displacement and the output voltage.

With signals which are stable to some degree, it is not necessary to scan every signal period. A number of periods can be omitted between scans. To this end, the sampling discriminator combines a harmonic generator and a phase discriminator, which is why it is often used with frequency multiplication in PLL circuits.

At low frequencies, the switch of the sampling mixer can make use of FET's, while Schottky diode bridges are used at high frequencies, up to the high gigaHertz range. The controlling pulse generator can be constructed for relatively long pulses (> 5nsec.) with logic gates, with efficient utilisation of the gate running time, while the step-recovery diode is the appropriate component for short pulses. The avalanche transistors, which were frequently used at one time, have now rather been forgotten. Fig.2 shows a section from the wiring diagram of the DB1NV 007 spectrum analyser, as an example of a sampling discriminator with a field effect transistor



DB1NV **Fig.3: I/P and O/P signals from Gates of Phase Discriminator**

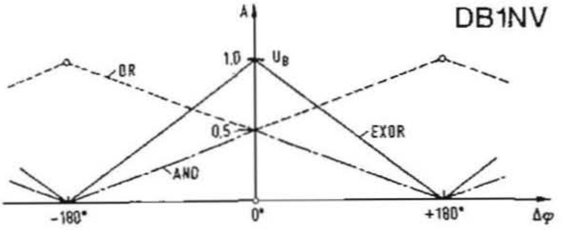


Fig.4: Phase Discriminator Characteristics of three Gate Phase Discriminators, showing DC Fraction of Output Voltage

switch and pulse generation using the gate cycle time. The reference frequency is 15 kHz, with an input frequency of between 7 and 20 MHz.

If the input signals are already present as square wave voltages, e.g. as output signals from frequency dividers, then digital phase discriminators are used, which can be integrated, together with the frequency distributors, as "one-chip PLL's".

5.3. Digital mixers as phase discriminators

All types of gate can be used as digital mixers, provided the input signals have a pulse-width repetition rate of 50%. Depending on the phase position of the input voltages, a square wave signal is obtained at the output with a variable pulse-width repetition rate; in other words, the average DC voltage at the gate output is a measure of the phase difference. Fig.3 (from (1)) clarifies the relationships between the three types of gate, AND, OR and EXOR ("Exclusive or"). In practise, the EXOR gate is preferred, because it supplies an output signal with double the frequency of the input signals. This simplifies the filtering of the output voltage. Moreover, the EXOR gates have the steepest characteristic line, as shown by Fig.4 (from (1)).

However simple the gate phase discriminator may appear, it still has a few disadvantages, which mean it is interesting only for the applications listed below:

- → The input signals must have a pulse-width repetition rate of 50%, so that the discriminator operates correctly
- → The output signal is a steep-sided rectangle, which means that the DC fraction can be separated out only by spending a lot of money on filters.

This discriminator is therefore predominantly used in low-frequency PLL's such

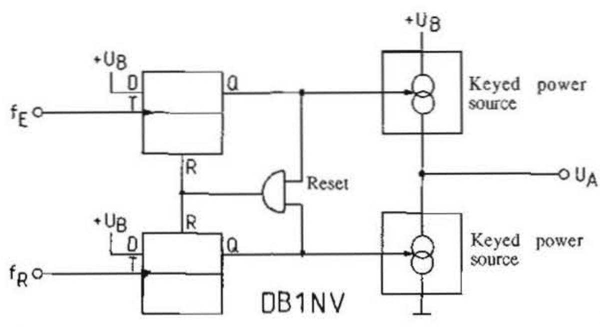


Fig.5: Basic Circuit of a Phase Frequency Discriminator

as modem circuits and the like.

All the phase discriminators listed so far have the following advantages and disadvantages:

→ Since the phase discriminator characteristics are symmetrical (Fig.4), it is not essential to ensure that the control potential has the correct polarity. Phase reversal in the variable-gain amplifier is of no significance.

→ At least one input signal must be (almost) symmetrical. Short pulses can not be processed.

→ If the frequencies are unequal, the differential frequency of the input signals appears at the output. If it is in the band-pass width of the loop filter, then the VCO is "warbled through" from this frequency until the PLL has synchronised. Under certain circumstances an "interceptor oscillator" has to support this process.

→ A remedy can be provided by a type of phase discriminator which reacts to only one flank of the signals and which also acts as a frequency discriminator in the unsynchronised condition, i.e. it leads the VCO to the scanning point without detours.

5.4. The phase frequency comparator

In its simplest form, the phase frequency comparator consists of two D flip-flops, a reset gate and two keyed power sources, as per Fig.5.. If both flip-flops react to the rising flank, the mode of operation is as follows:

If the positive flank of input 1 sets the associated flip-flop, the power source is activated, which pulls the output in a positive direction. If the positive flank of input 2 reaches the associated flip-flop, then it is likewise set and the power source is activated, which pulls in a negative direction. This condition lasts only a short time, as the reset gate recognises both set flip-flops and resets them both simultaneously. The output current of the circuit is thus a positive pulse, the width of which corresponds to the temporal interval between the input signal flanks. As you will easily see, when the phase displacement at the output is reversed, a negative current pulse arises.

The behaviour of the circuit with signals with different frequencies is interesting. The signal with the higher frequency always switches on its flip-flop until the slower signal triggers a reset and resets both signals. Thus only the faster signal generates output current pulses, which corresponds to the frequency discriminator function desired.

With balanced signals, on the other hand, in general no output pulse is generated, if we ignore the brief interference spikes when the flip-flop is reset. In the locked condition, its output signal is a pure DC voltage, so that a simple capacitor at the output is sufficient to act, now and again, as a loop filter.

The phase frequency comparator is therefore often used in integrated PLL circuits.

However, this circuit has a disadvantage, which admittedly is often insignificant, but explains many noticeable effects in a frequency synthesiser. If both flanks meet the inputs almost simultaneously, then the width of the output pulse is no longer dependent on the flank interval, but only on the signal running times in the flip-flops and in the reset gate. There is an area around phase coincidence in which no output voltage proportional to the phase position is generated. Specialists in automatic control technique speak of a "dead zone". If the PLL loop is locked, then the VCO can drift off a little, until the phase difference is big enough to generate a control potential which compensates for the drift again. This is noticeable in that, in synthesisers of this type, the VCO continuously "flutters" around its rated frequency, which is particularly striking when an SSB receiver is used for listening.

In FM radio equipment, this low-frequency spurious modulation is mostly negligible. There are three possible remedies, which represent different compromises in terms of expenditure and success:

1. The logic of the discriminator is made as fast as is possible with the circuit technology used. This reduces the dead zone. Old synthesiser IC's using PMOS technology, such as the S187 from Siemens, are particularly unfavourable in this context. Rapid logic naturally increases the current consumption.
2. The phase frequency comparator is used only for interception and for rough synchronisation. An analogue phase discriminator (usually a sampling discriminator) then takes over the fine control. This technique is used in the NJ8820 from Plessey and, in a similar form, in the MC145159 from Motorola. The results are optimal for what is, of course, considerable expense.
3. Two phase frequency comparators are used, which are powered by signals phase-displaced through gate running times, so that only one discriminator is ever in the dead zone. The output pulses are suitably combined in the IC. The phase discriminator in the TBB200 from Siemens should operate by means of this procedure.

It can be seen from the above that the selection of a suitable phase discriminator plays a key role in a successful synthesiser project. True, nowadays most tasks can be mastered using one-chip synthesisers, which contain the entire digital logic of the reference divider, adjustment divider and phase discriminator. But frequently only a discretely constructed phase comparator is in a position to guarantee the signal quality required.

A similar critical element is represented by

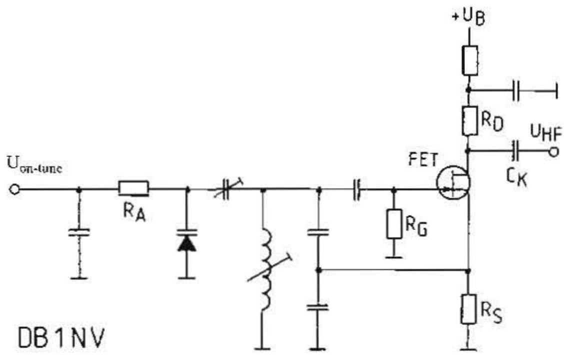


Fig.6: Typical VCO Circuit with FET

the voltage-controlled oscillator or VCO, which is the subject of the next section.

6. VOLTAGE-CONTROLLED OSCILLATORS

The VCO's (voltage-controlled oscillators) used in PLL circuits can be roughly divided into RC circuits and LC circuits. RC oscillators, usually multi-vibrator circuits, are used in the low-frequency range, since they are easy to lay out for low oscillation frequencies. If the requirements for stability are not high, they can also be used in the lower part of the high-frequency range. The cycle recuperation PLL in floppy and hard disk controllers represents one common application.

LC oscillators with capacitance diode tuning are used, almost exclusively, in the high-frequency range. A typical circuit of this type is the FET oscillator in Fig.6. The structure will be recognised as that of a Colpitts oscillator with feedback from source to gate. The HF is tapped at the drain of the FET with low feedback. A capacitance diode connected through a capacitor displaces the frequency as a function of the control voltage. However, this basic circuit has its peculiarities, and high-quality VCOs can not be constructed unless they are known. This basic circuit will later serve us as a model for the explanation of crucial construction guidelines.

"PLL experts" often express the opinion that there is no need to pay much attention to stability and drift in the construction of a VCO, since the oscillator is actually connected to the reference crystal and should be allowed to drift and generate background noise. We already know from the first part of this series of articles that, in principle, only those VCO malfunctions which lie within the band width of the control loop are controlled, and this is often only a few hundred Hertz. It is thus a good idea to construct the VCO as solidly as a good VFO. If, for example, the VCO in a radio set is tuned with an externally fed-in, stable voltage, with the frequency remaining on the reception channel for minutes without readjustment and no noticeable hum modulation or howl back, then the circuit may be considered a success.

6.1. Critical points

The following points are to be taken into account in the circuit design and compromises are to be made:

1. Basically the quality of the frequency-determining oscillation circuit or resonator should be selected to be as high as possible, to keep the static sidebands as narrow as possible. Narrow static sidebands mean a high short-time stability for the oscillator. The influence of the quality can be explained by saying that the unavoidable static from the oscillator transistor due to the band tuning effect of the resonator is reduced to a narrow range near the carrier. To ensure that the transistor does not work against a high unloaded Q for the oscillation circuit, it is loosely coupled to the circuit.
2. Capacitance diodes have relatively good qualities in the VHF and

UHF range. If a wide tuning range is required, then the capacitance diode must be permanently connected to the circuit, which impairs the quality and the static characteristics. For high-quality oscillators, there is the possibility of splitting the entire tuning range into several sub-ranges. The most reliable solution is to use separate oscillators for each sub-range.

3. If a continuously tuneable oscillator with an extremely wide tuning range is required, then the poor static characteristics must be accepted. Since in this case the resonator quality is determined exclusively by the capacitance diode(s), it makes no sense to invest considerable mechanical resources in a high-quality resonator. A neatly constructed coaxial circuit with strongly linked diodes does not bring about better characteristics than a micro-strip layout.
4. At first sight, the long-term stability of the VCO seems to have no significance, since it is actually balanced by the control loop. This is also a false conclusion, since an oscillator which is very dependent on temperature must be able to offer a large "tuning reserve" on both sides of the useful tuning range, so that the PLL circuit can still intercept the oscillator at extreme temperatures. But there again, an unnecessarily wide tuning range is costly in terms of signal-noise ratio.
5. Only bipolar types and barrier layer FETs, both of which are characterised by both good high-frequency characteristics in the low-frequency range and low static, come into consideration as oscillator transistors. MOS and MESFETs generate a strong flicker effect in the low-frequency range and the lower part of the high-frequency range, which is immediately transferred to the oscillator signal as AM modulation. Since the input and output capacities of a transistor are independent of level control, the amplitude modulation is converted in the transistor into a frequency or phase modulation, which can no longer be removed by subsequent measures such as, for example, using a limiter.

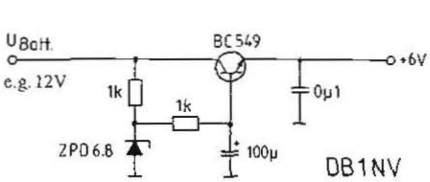


Fig.7: Low-noise simple Voltage Controller

7. CIRCUIT DETAILS

Once the basic dimensions and the selection of components for a VCO has been clarified, it is recommended that you take a look at a few circuit details which are not obvious, but which can have a decisive influence on the quality of the oscillator. Some of these things were already being discussed years ago in the amateur literature. But if we look at the failed designs of some recent synthesisers we can see that many things have been forgotten again.

7.1. The operating voltage

Every oscillator reacts to oscillating power voltages with a frequency modulation, since the transistor capacities depend on the voltage. The extent of the modulation depends both on the resonator quality (high quality reduces the frequency fluctuation) and on the coupling of the voltage-dependent transistor capacitors to the oscillation circuit.

It is thus reasonable to allocate the oscillator its own voltage regulator. But before you reach into the box with the integrated tripod voltage regulators, there's one thing you should think about. The worst interference comes from rapid oscillations of the supply voltage in the form of superimposed high frequency and static, since these can be only imperfectly controlled by the PLL, or not at all. Tripod regulators have poor control characteristics at high frequencies and themselves produce some background noise. The "old" regulators in the 78XX range are relatively good here. Modern low-drop regulators and energy-saving low-power layouts exert a considerably slower control and also produce a lot of noise. The best solution is a discretely constructed regulator with a Z diode and an emitter follower. Sufficient current should be supplied to the Z diode and an RC member should be mounted between the diode and the base, as per Fig.7, in order to suppress the diode static.

7.2. The bias voltage supply to the tuning diode

Many oscillator circuits can be found in which the bias voltage for the tuning diodes is fed in through high ohmic resistances (100kΩ to 1MΩ). This solution has the advantage that a resistance is cheaper than a choke and moreover oscillator malfunctions caused by natural choke resonances can be avoided. But every diode driven in the high-resistance direction has an inverse current which undergoes statistical oscillations, and thus represents a DC current with a superimposed AC static current.

At a high-ohmic resistance in the DC circuit of the diode, this static current generates a noise voltage, which is superimposed on the tuning voltage and is audible as a noise modulation. So for high-quality VCO's it makes sense to design the DC current path of the tuning voltage to be low-ohmic. The best solution is attained if the diode voltage is fed in through already existing coils, as in the oscillator for ceramic resonators described by the author in VHF Communications 4/90. Should this not be possible for reasons of circuit technology, a series circuit, consisting of a choke and a damping resistance of approximately 1k, should be used instead of the series resistance. Dimension the choke for about 10 or 20 times the oscillation circuit inductivity. Should oscillator malfunctions arise, such as breaking down of the oscillation at certain frequencies or "reverse tuning", then the natural choke resonance is within the variable frequency range, in spite of the damping. A different choke inductivity solves the problem.

For simple oscillators, e.g. for walkie-talkies, feeding the voltage in through a resistance is thoroughly acceptable, provided it is made only as high-ohmic as is absolutely necessary. Usually even 5 to 10kΩ are sufficient from the point of view of possible damping of the oscillator circuit.

The fact that the noise characteristics of a VCO and thus of the synthesiser can be decisively improved by correct operational voltage filtering and the feeding in of a tuning voltage was already being written about in articles in CQ-DL and VHF Communications some years ago. However, many younger circuit developers seem to have lost this knowledge, as is shown by negative examples from recent times.

7.3. The low-frequency wiring of the oscillator transistor

The oscillator transistor amplifies not only the wanted signal but also its background noise, so that the noise spectrum extends from the lowest low-frequency range (so-called flicker effect) to the limiting frequency of the transistor. This noise voltage now modulates the transistor characteristics like an external interference voltage, in particular the natural capacitancies, and appears as a noise modulation on the output signal. The width of the static sidebands of the output signal depends on the resonator quality of the oscillator, since the resonance circuit is responsible for the band pass filtering of the output signal. How strong the effect of the low-frequency noise from the transistor is depends on the circuitry on the low-frequency side, i.e. on the low-frequency amplification of the transistor. If the transistor is operating with full open-loop voltage gain, then the noise level at the collector will cause a correspondingly powerful noise modulation.

Thus the goal must be to reduce the low-frequency amplification of the transistor by negative feedback, and to design the low-frequency source resistance in such a way that the transistor is operated with its optimal source impedance for the low-frequency noise.

One possible solution for bipolar oscillators has been described in the VCO's DB1NV 012/013. Here a collector resistance more effective for low frequencies, in conjunction with low-frequency feedback from the collector to the base, provides for good phase noise characteristics. The disadvantage of such feedback is that it extends the time between the switching on of the oscillator and the reaching of the operating point, which should be taken into account for keyed oscillators (transmitter / receiver circuit).

For FET oscillators like the one in our example (Fig.6), the gate is often closed at high resistance, so that a negative bias can be built up by rectifying the high frequency and the operating point can be stabilised. The noise voltages arising at this resistance can be rendered ineffective in that the gate

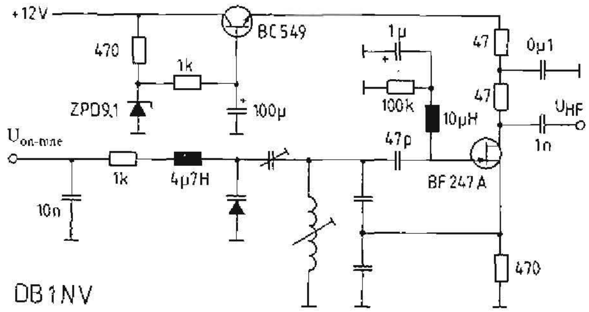


Fig.8: Noise-optimised VCO

is put high up, using a high-frequency choke, and the gate leakage resistance is bridged by a capacitor.

The changes just described can also be seen in Fig.8, which shows our specimen VCO after noise optimisation.

8. OTHER COMPONENTS OF A SYNTHESISER

The other components of a PLL synthesiser, in addition to the VCO and the phase discriminator, are the reference oscillator, reference divider, adjustment divider and variable-gain amplifier with low pass filter.

The existing functional units in the PLL IC are normally used for the dividers in the signal and reference branch, or else recourse is had to logic IC's using ECL or HCMOS technology, so no further explanation is necessary. The reference oscillator is also integrated in most PLL IC's, so that if high-quality crystals are used reasonable oscillator characteristics are obtained straight away.

If the requirements for frequency stability are more stringent, the use of an external temperature-compensated crystal oscillator (TCXO) is recommended. Since TCXO's with the standard frequencies of 6.4 MHz or 12.8 MHz are a constituent part of every mobile telephone and are thus a mass-produced article, you don't have to pay much more for them than for a made-to-order crystal. Prices start at £20 and rise in accordance with the stability requirements and temperature range.

Those wishing to construct their own reference oscillators should in any case dispense with TTL gate oscillators, such as can be found in many old publications. Even in application documents for microcomputers, there have since been warnings concerning the instability and tolerance sensitivity of circuits of this nature.

The variable-gain amplifier and low pass filter functional units are of essential importance in keeping the quantity of spurious emissions low and for the transient response of a synthesiser, for which reason we can not avoid individual dimensioning. The author has no wish to bore his readers with the mathematical description of a PLL circuit, and prefers instead to give clear assistance in dimensioning. Anyone interested in the mathematical description can consult, for example, (2).

The variable-gain amplifier with a low pass has to fulfil the following functions:

→ Amplifying the output voltage of the phase discriminator to a value sufficient to control the capacitance diode in the VCO.

→ Filtering out the residual ripple of the phase detector signal, in order to prevent any interference modulation of the VCO through the remains of the phase monitoring frequency.

→ Ensuring the stability of the control loop using suitable filter characteristics.

The last two points usually have contradictory effects on dimensions, so you have to look for compromises or have recourse to expensive filter structures. Now how should the low pass filter be dimensioned so that the control loop operates in a stable manner, i.e. avoiding a malfunction and without "resonant rise"?

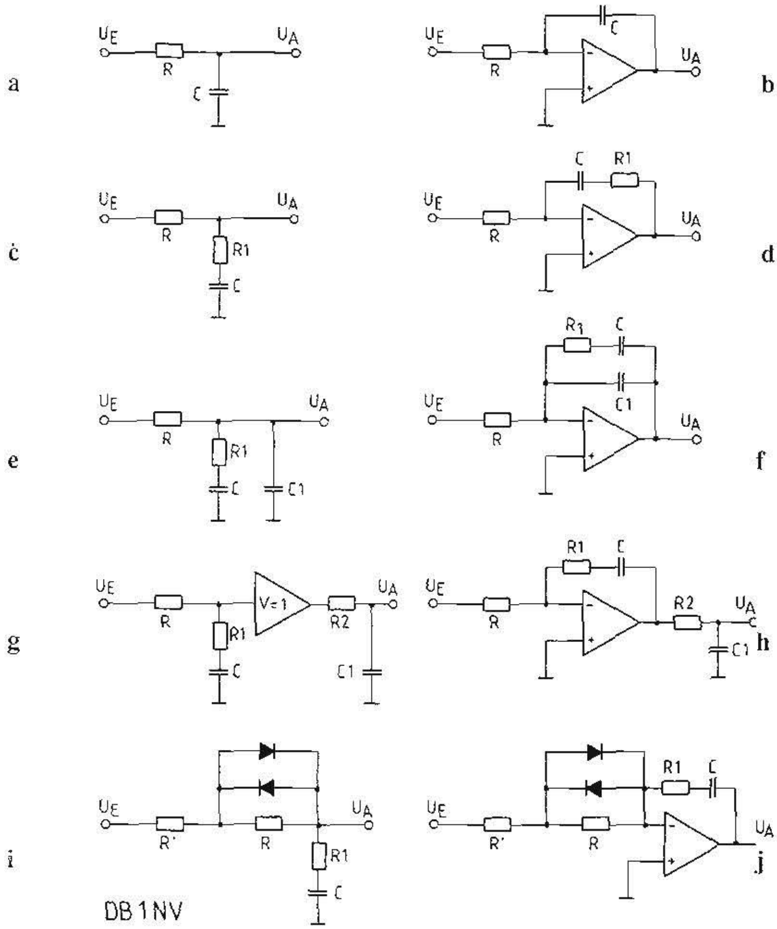


Fig.9: Loop filter options for PLL circuits

A PLL is nothing more than a feedback amplifier, which keeps the VCO frequency steady against external and internal interference factors. If all components are ideal and free from delay, then no low pass is needed in the loop. But since all functional units have an operating delay, the feedback produces a spurious feedback at a certain interference frequency, so that the interference builds up. The low pass now reduces the amplification in the control loop for high frequencies to such an extent that the feedback condition is no longer fulfilled for the critical frequency.

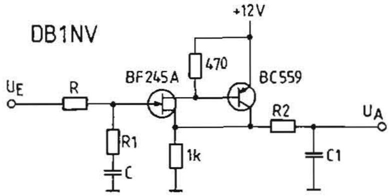
The setting divider and the phase discriminator make the biggest contribution to this delay time, as can easily be imagined. If the input frequency of the divider alters, then the phase discriminator does not "notice" this until a signal flank has arisen at the output of the divider. The discriminator itself needs one flank each in the signal and reference branches to determine the phase differential.

It can thus be seen that the maximum sensible control speed, and thus the limiting frequency of the low pass filter, can amount to only a fraction of the phase monitoring frequency.

If, as a first step, the low pass is dimensioned at 1/10 to 1/50 of the phase monitoring frequency, then the PLL can at least be engaged and can then be optimised.

8.1. Various low pass filters

In constructing the low pass filter, you have the choice between passive filters (the low pass is mounted in front of or behind the variable-gain amplifier) and active filters, in which the low pass is integrated into the feedback of the variable-gain amplifier. Figs. 9a to 9h give a summary of common filter structures. In Fig.9a, we see a simple passive loop filter, a first-grade low pass with the limiting frequency f_g = 2.R.C. The active version in Fig.9b is an integrator with the time constant t = R.C.. The difference between the two options is that with the passive filter the phase difference between the reference signal and the VCO signal is not set to zero, but a phase error remains which depends on the frequency, because the output voltage of the phase discriminator is the tuning voltage of the VCO. The active loop filter integrates the phase error at zero and keeps it there (or at a fixed value conditioned by the circuit). The active loop filter can be used with advantage provided the output voltage of the phase discriminator does not allow any sizeable deviation from its equilibrium value. The simple first-grade loop filter can not provide an optimal transient effect during interference or frequency changes.



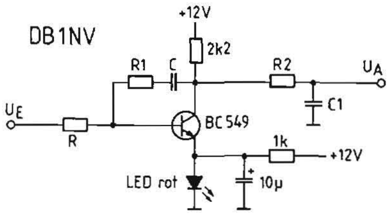


Fig.10: Examples of non-inverting (a) or (b) variable-gain amplifiers

As stated earlier, every PLL has an internal delay time, which generates a phase angle rotation in the control circuit and converts the feedback into a spurious feedback. The first-grade loop filter is already generating a phase angle rotation of 45° at its limiting frequency, which increases to nearly 90° at still higher frequencies, and is added on to the phase angle rotation of the rest of the loop. Thus the filter at first just impairs the "phase reserve" of the PLL loop and the limiting frequency must be selected lower than the optimum for reliable operation.

If a resistance, R1, with a value of app. 1/10 R, is wired up in series with the filter capacitor, C, as per Fig.9c or 9d, the transient effect of the PLL after interferences or frequency changes improves decisively. In accordance with what has just been said, the effect of the extra resistance can easily be understood. As the frequency increases, the low pass reduces the amplification in the control loop, but the interference phase angle rotation of the filter is much slighter than before. At high frequencies, the capacitor can be considered as a short-circuit, and we obtain a voltage divider made up of R and R1, which causes no phase angle rotation.

Unfortunately, this loop filter has a disadvantage. Whilst damping using the filter as per Fig.9a increases with the frequency, by 6 dB / octave, with the filter as per Fig.9c it reaches a constant value, namely the voltage divider ratio from R and R1. Since the output of a phase discriminator always contains residues of the phase monitoring frequency and its harmonics, these fractions are better damped by the circuit as per Fig.9a than by that as per Fig.9c. A modulation of the VCO with these interference voltages should be unconditionally avoided, since they are expressed as spurious emission symmetrical to the rated frequency at intervals corresponding to the phase monitoring frequency and its multiples. In the extreme case, this gives the notorious "synthesiser lattice fencing", which seriously interferes with the adjacent channels. One remedy is provided by an additional low pass, which becomes effective only from the phase monitoring frequency onwards, and has only an insignificant influence at lower frequencies. The low pass can be supplemented by a capacitor, C1, as in Fig.9e or 9f, which should be dimensioned at about 1/10 C. R and C1 then form the additional filter for the phase monitoring frequency and its harmonics. This circuit is also often described as "two time-constant filters".

A further improvement can be obtained if the second low pass is mounted behind the variable-gain amplifier, as shown in Fig.9g and Fig.9h. In the first place, the low pass for filtering the phase monitoring frequency can here be dimensioned independently of the rest of the circuit, and in the second place it also attenuates high-frequency interference factors stemming from the variable-gain amplifier, such as noise.

For synthesisers with a rapid transient time after a frequency change, a further improvement is possible using a non-linear loop filter as per Fig.9i or 9j. One section of the longitudinal resistance, R, is bridged over with two anti-parallel diodes. For large frequency jumps, which entail a big change in the output frequency of the phase discriminator, the capacitors of the loop filter are rapidly reversed, so that the diodes become high-ohmic again, and the optimal time constant for static operation is selected.

The pre-requisite for the usability of a non-linear loop filter is that the phase discriminator delivers a DC voltage which is not interspersed with spike pulses. The spikes would open the diodes and would arrive at the filter output without much weakening. In this respect, a sampling phase discriminator is ideal.

8.2. The variable-gain amplifier

In the selection of the variable-gain amplifier, it should be borne in mind that its noise goes directly into the VCO and becomes noticeable as phase jitter. The best solution is usually a discrete amplifier or impedance converter with one or two transistors - some examples are shown in Fig.10. Operational amplifiers are easier to handle, but have to have low noise levels. "Ancient types", such as the "741" or the CMOS-OP's, such as, for example, the TLC 27M7 manufactured by Texas Instruments, are completely unsuitable. CMOS and micro-power operational amplifiers make more noise than bipolar types or those with a FET input. The author has had some good experiences with FET types such as, for example, the TL070 family from Texas Instruments, or special low-noise types, such as the OP27 from PMI or Analog Devices. The best price-performance ratio, in the author's opinion, is provided by the NE5534 from Philips, which was specially created for high-quality low-frequency applications, e.g. in studio technology. Its small disadvantage, the high input static current, can be eliminated if necessary by a series-connected source follower with standard FET's, such as the BF 245.

9. CALIBRATING AND OPTIMISING THE CONTROL LOOP

9.1. Optimising on a mathematical model

To optimise the low pass filter and thus the control behaviour of the PLL control loop, we can analyse the circuit mathematically and then hope everything works. To prepare the calculation, we have to determine the control gradients of the phase discriminator and the VCO, which requires a certain amount of measurement. The low pass must then be calculated either "on foot" or using a suitable computer program. Since the effort involved is considerable, in particular for people little inclined to mathematics, the author suggests the following experimental procedure:

9.2. Optimising through experimenting

The point of departure is a loop filter as per Fig.9c or 9d, for which the following dimensioning rules have proved themselves with normal radio synthesisers with 25 kHz channel grids, which then serve as a point of departure for the optimisation. With a simple receiver-synthesiser, the time constant RC will initially be selected as a millisecond (e.g. R = 10kΩ, C = 0.1F), and the resistance, R1, will take the form of a trimming potentiometer, with a value of 1/2R.

If the synthesiser is to be modulated in the FM transmitter, then the RC time constant should be selected at approximately 10 to 50 milliseconds, since a PLL which is too rapid will otherwise level the low-frequency fractions of the modulation.

With these dimensions, the PLL circuit should engage when R1 overflows, i.e. the control voltage of the VCO is a pure DC voltage, which alters when the channel changes or when the VCO is manually de-tuned. If the control voltage "sticks" at a stop, there may be a number of reasons for this:

The tuning range of the VCO is insufficient to reach the rated frequency.

The control voltage is unintentionally being inverted, so that the frequency discriminator, integrated in many PLL IC's, is pulling in the wrong direction. Depending on the type of IC, a passive, non-inverting or active filter is pre-supposed, and the control direction can often be reversed in the IC!

With PLL circuits with ECL pre-dividers, the error may lie in too little controlling of the divider by the VCO. If there is too little control, or no control, many pre-dividers oscillate wildly and the control circuit runs permanently, trying to stabilise the situation.

If control oscillations occur which do not disappear when R1 is altered, then initially the filter capacitor is enlarged until the circuit operates in a stable manner. With active loop filters, the problem may also lie in an oscillating operational amplifier.

Once the synthesiser is operating to this extent, it can be established, by listening to the adjacent channels, or (if possible) by using a spectrum analyser, whether spurious emissions at the phase monitoring frequency are present. Next, it should be decided whether an additional filter is needed, as per Figs. 9e to 9h. The dimensions applying here are: time constant R.C1 / phase monitoring frequency. If this does not bring the spurious emissions under control, then check whether the static generally is coming into the VCO through the control voltage, as opposed to digital pulses finding their way directly into the VCO.

9.3. Optimising the control behaviour

If the synthesiser has operated satisfactorily up to now, we can go on to the optimising of the control behaviour. To this end, periodic interference with the control loop is required, the stabilising of which is observed using an oscilloscope, which is connected to the tuning voltage of the VCO. The control procedure should run briskly and without overshooting. If enough is known about computers, the synthesiser IC can be moved up and down a few channels at a frequency of between 10 and 20Hz. Otherwise, a square-wave signal with a similar frequency is connected into the control loop, as per Fig.11.

The R1 and C values of the low pass filter are optimised in such a manner that the square-wave interference is stabilised as fast as possible and without oscillations. With a synthesiser which can be modulated, the transient load response is dimensioned to values of app. 20 milliseconds. The calibration is advantageously carried out at a frequency at the lower end of the VCO tuning range, since here the tuning gradient and thus the risk of instabilities is at its greatest.

At the top end of the band, then, there is strongly damped control behaviour, and thus a longer response time. This creates less interference than oscillations at the lower end of the band. Anyone wishing to perfect the control behaviour can use phase discriminators with programmable control gradients and select the appropriate control gradient in each case on the basis of the frequency selected.

It will be noted that freedom from spurious emissions and the desired control behaviour can often be brought about only with compromises on a denominator. Assigning low-frequency dimensions to the low pass for the phase monitoring frequency (C1)

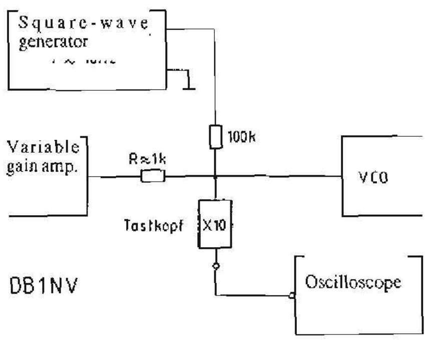


Fig.11: Measurement circuit for optimising PLL

increases the spurious emissions interval, but increases the over-shots during the transient effect.

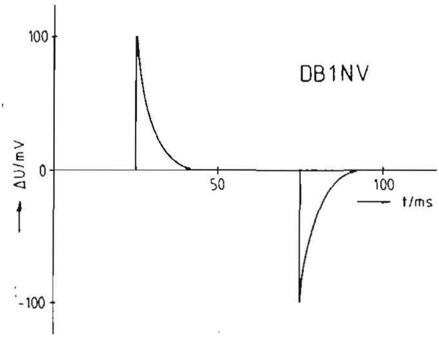
Fig's. 12a and 12b show the control behaviour of a correctly dimensioned PLL and an insufficiently damped PLL, as obtained using the Fig. 11 measurement circuit.

10. FULLY SOLID-STATE PLL FUNCTIONAL UNIT

Until a few years ago it was still normal to assemble the digital part of a synthesiser from individual logic IC's. Since then, many one-chip PLL circuits have become available, which contain all the digital elements, from the setting divider through the phase discriminator, the reference oscillator and the reference divider to the control logic for the dual mode pre-divider. Frequently, we even find a section of the variable-gain amplifier or a loading pump for the generation of a higher VCO tuning voltage in the IC's. Since a micro-controller controls the operating sequence in almost all radio equipment, the directly programmable synthesiser IC's are being squeezed out by layouts with bus connections.

The most common types of interface are:

- 4 bit parallel with additional address inputs for direct connection to the data buses of 4-bit and 8-bit micro-controllers
- Three-wire bus, a serial synchronous interface, which consists of a data line, a timing line and a reception line; this bus is often described as a "microwire"



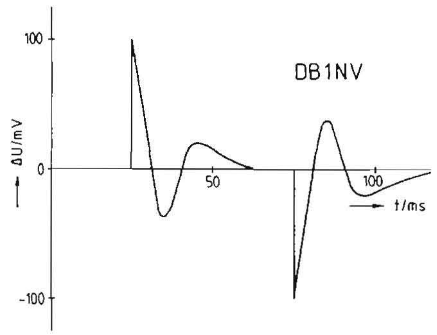


Fig.12: Control behaviour of correctly dimensioned (a) and insufficiently damped (b) PLLs

- The I^2C bus, known from entertainment electronics, which consists of only a data line and a timing line

Anyone interested in the data transmission procedure should refer to the manufacturers' special data books (3) - (8). The leading manufacturers of synthesiser IC's are Fujitsu, Motorola, Plessey and Siemens.

The author has restricted himself to products from the Motorola and Plessey companies, since here the distributors (Enatechnik for Motorola, Astronic for Plessey) willingly made small amounts of equipment available for experiments. Unfortunately, many interesting components exist only in the "handy 2500-unit big pack", which makes them unsuitable for amateur use.

10.1. "Wanted list" of selected PLL modules

The following data summary may provide the interested reader with a survey of the data and operational experience relating to some selected PLL components:

MC145152 (Motorola):

This CMOS synthesiser, already decidedly elderly, can be parallel-programmed using 19 data lines, but subsequently requires no micro-processor for control. It operates with all common dual-modulus pre-dividers from Motorola and Plessey. The disadvantage is that the divider factor for the reference frequency is programmable only to 8 places and the phase discriminator has two separate digital outputs which are merged externally in the variable-gain amplifier. Since the phase discriminator generates pronounced residual pulses in the engaged condition, considerable work has to be spent on the filter in order to bring the spurious emissions under control.

MC1451146 (Motorola):

The MC145146 is designed for programming by means of a 4-bit data bus and, as well as the MC145152 equipment, possesses a fully programmable reference divider and a second phase discriminator, which avoids the "dead zone" mentioned in connection with simple phase frequency discriminators.

MC145159 (Motorola):

The MC145159 is programmed by means of a three-wire bus and has similar characteristics to the IC's above. However, it has been optimised for use in high-quality synthesisers. So it has an analogue sampling phase discriminator, which supplies a very clean output voltage to the succeeding units, without digital interference pulses. Separate supply and earth connections for logic and for the analogue section make it possible to provide the phase discriminator with a clean operating voltage. A digital course discriminator is responsible for the rapid engagement of the PLL. The MC145159 is laid out for operation with a passive, non-inverted loop filter.

NJ8820 (Plessey):

The NJ8820 is especially suitable for high-quality synthesisers with relatively few operating channels, for it possesses a special "PROM interface", and can read its program data off independently from an EPROM. The reading procedure is triggered using an input which is connected only by capacitive coupling to the PTT line. Each level change then automatically acts as a re-programming. The NJ8820 has a digital course discriminator and a high-quality sampling discriminator for fine control. An active, inverted low pass filter is needed. The NJ8821, with a 4-bit parallel bus, and the JN88C22, with a serial wire bus, also belong to the same family of circuits. All the IC's have been produced using CMOS technology and require external dual-modulus pre-dividers for frequencies greater than 10 MHz.

SP8853 (Plessey):

The SP8853, laid out in accordance with bipolar technology, does have a relatively high current consumption but, to make up for that, it has a dual-modulus pre-divider up to 1.5 GHz, so that it can be described as a genuine one-chip solution. The SP8853 is programmed by means of a three-wire bus. More rapid frequency changes are possible without re-programming, thanks to a doubled frequency register, which is selected using a switching input. Coarse and fine phase discriminators with programmable output currents and a programmable polarity for the output voltage mean this module is pre-destined for professional applications, which is reflected in the price. An identical format is available in a cheaper plastic housing under the description SP8861.

10.2. "Wanted list" of selected pre-dividers

In most cases, a matching dual-modulus pre-divider is required for the synthesiser IC's referred to above. The following types from the very wide Plessey range were investigated:

SP8716-19:
This family of dividers typically operates at up to 520 MHz and has a typical current consumption of 7 - 10 mA. The SP8716 divides by 40 and 41, the SP8718 by 64 and 65, and the SP8719 has divider factors of 80 and 81.

SP8703/04:
The SP8703 and its low current-drain successor the SP8704 are designed for input frequencies of 1 GHz. The current consumption of the SP8703 amounts to 30 mA and it can be switched off with a power down input. The divider factor is 128 or 129. The SP8704 requires only 10 mA more and can be wired for a 64/65 or 128/129 division.

SP8793:
The SP8793 divides by 40 or 41 and operates at up to 225 MHz, with a typical current consumption of 4 mA. An internal voltage controller makes operation possible at 6.8 to 9.5 V, if no 5 V supply is available.

It should be mentioned here that Fujitsu, Motorola and Siemens also supply interesting pre-dividers, but they are difficult for the ordinary amateur to acquire. Perhaps a distributor interested in amateur radio may have the modules available.

Following this comprehensive description of the modules making up a frequency synthesiser, dear reader, you should be able to design and make a working synthesiser to suit your own requirements. Perhaps you have also experienced one or two sudden insights which now explain earlier false steps in synthesiser engineering.

The third article in the series will deal with the description of two rather more complex synthesisers.

11. LITERATURE

(1) Arnoldt, Michael: Standard Frequency and Time Signal Reception

(2) Walz: PLL Engineering Franzis-Verlag, 1989

(3) Fujitsu Telecommunication Products 1992 Data Manual

(4) Motorola CMOS Data Manual, Volume 2, Special Functions, 1983

(5) Motorola MECL Device Data, 1989

(6) Plessey Professional Products, May, 1991

(7) Plessey Personal Communications Handbook, June, 1990

(8) Siemens IC's for Industrial Applications, 1990

Wolfgang Schneider, DJ 8 ES

CW Call Sign Transmitter

Every radio amateur has to carry out test transmissions now and then. Whether it be for test purposes (TVI/ BCI test) or for comparison, the question always arises: how do I modulate my test signal?

The following description introduces a simple and yet efficient CW call sign transmitter. Dashes, dots, call signs or even short texts can be transmitted in CW mode by this assembly. In addition to rendering the operator's own test signal clearly identifiable, this also means that the regulations governing amateur radio are adhered to when the operator's own call sign is transmitted.

1. THE CW CALL SIGN TRANSMITTER

This CW call sign transmitter puts a small and at the same time cheap accessory into

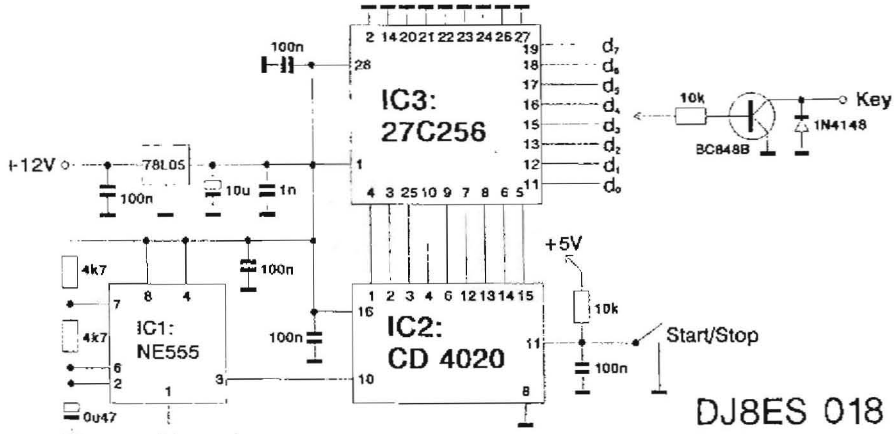


Fig.1: The CW Call Sign Transmitter

the hands of the user. A circuit has been created here at minimum cost which meets all the necessary requirements. The following programming should be taken as standard:

- → One line of dots
- → One line of dashes
- → One line of dots and dashes
- → Test loop (TEST DE CALL)
- → Test loop (CALL LOCATION, continuous dash)
- → cq loop (CQ CQ CQ DE CALL CALL CALL PSE K, break)

Of course, the CW call sign transmitter can be programmed in accordance with personal ideas and wishes. In the circuit shown, the assembly is restricted to 512 storage steps per text. This storage volume corresponds to approximately one minute's text at a rate of 60 bpm.

The complete memory area can be made use of through suitable modification (address selection of EPROM). This corresponds to a maximum of 32 k of CW text.

2. DESCRIPTION OF CIRCUIT

A type 27C256 (IC3) EPROM acts as a text memory and forms the core of the CW call sign transmitter. The storage chip is organised at 32 k x 8 bits. These 8 bits are used to store a maximum of any 8 CW texts.

The CW signs are read out serially. For this purpose, the EPROM is correspondingly controlled through its address lines. This function is handled by a CMOS-IC CD 4020 (IC2). The divider is a 14-digit binary meter, with the 9 low-value bits serving to control the EPROM.

The programmed CW texts should be transmitted at 60 bpm. With the memory area of 512 bits available for each of the eight possible texts (EPROM addresses 0dec - 511dec), the necessary clock frequency of 273 Hz is calculated.

A standard circuit with an NE555 timer IC

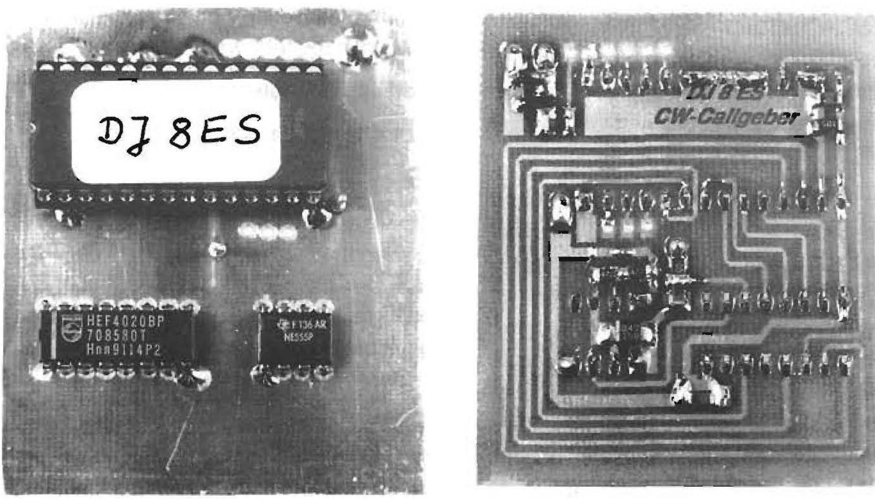
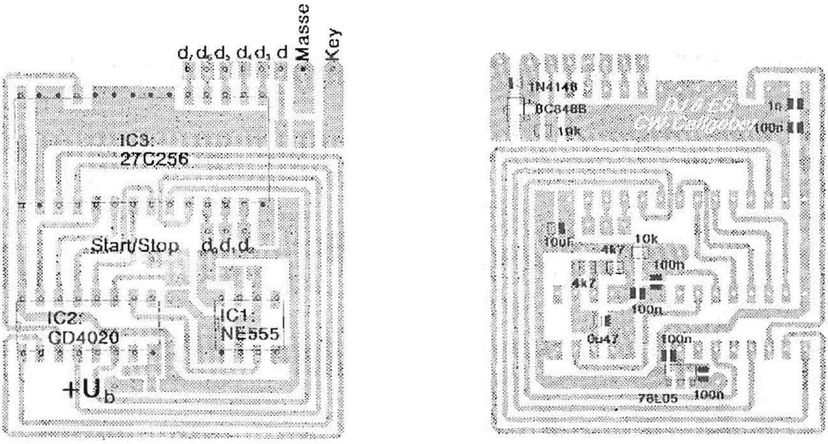


Fig.2: The prototype from the components side and the foil side



DJ8ES 018

DJ8ES 018

Fig.3: The circuit can be built with minimum expense

(IC1) is assembled as a clock. The output frequency is determined by the RC combination 2 x 4k7 and 047. To vary the speed, one of the resistances can be replaced by a potentiometer if necessary.

The circuit is served by a single-pole switch at the reset input of the binary meter (IC2). It distinguishes between "Start" (switch closed) and "Stop" (switch open). In the "Stop" position, IC2 is reset to address 0dec. The possibility of switch chatter keeps the RC unit out of 10k and 100nF.

Up to 8 different CW texts can be stored and called up using the given circuit. The desired EPROM output is selected using a step switch. To control the stages of transmission, or to control a transceiver, the switch output is provided with a BC848B transistor. A 78L05 voltage controller allows the call sign transmitter to be operated at the normal operating voltages (e.g. +12V). The current consumption of the assembly at 12V is only 7mA, so that it can be operated with a battery.

3. PROGRAMMING

The address area 0_{dec} - 511_{dec} is always available for the individual CW texts. The characters are serially distributed in the EPROM. A dot occupies 1 bit. If the programming initially proposed is used as a base, then the following system applies:

d0 = line of dots

d1 = line of dashes

...

It is important that address 0_{dec} should always remain empty (logic 0). When the switch is in the "Stop" position, the unit is reset to this address. This should not cause any character to be transmitted. The first CW sign starts with address 1_{dec}.

The EPROM is individually programmed using suitable aids (software, programmer). A sample listing can be obtained from the publishers.

4. ASSEMBLY INSTRUCTIONS

The CW call sign transmitter is assembled on a double-sided epoxy-coated board measuring 53 mm. x 72 mm.. The IC's are fitted from the components side and the SMD components from the solder side.

The board is small enough to be incorporated in a standard tinplate housing. The most varied assembly options are conceivable, e.g. step switches and start / stop switches also mounted inside the housing.

During assembly, make sure that the copper surfaces remain around the bores for all earth connections. They are soldered on both sides.

4.1. List of components

IC1	NE555 timer
IC2	CD4020 binary meter
IC3	27C256 EPROM

All other components as SMD layout:

1 x	TA78L05F voltage controller
1 x	BC848B transistor
1 x	1N4148 diode
1 x	10F / 20V tantalum electrolytic capacitor
1 x	0.47F / 35V tantalum electrolytic capacitor
1 x	1nF ceramic capacitor
5 x	100nF ceramic capacitor
2 x	4.7k
2 x	10k

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Crystal Testing

Oscillator crystals are nothing new. Technical applications for their piezo-electric effects have been known for 60 or 70 years. Their characteristics have been known about for just as long. Manufacture was and is constantly refined. Nevertheless, a crystal is still a relatively expensive component. And you don't throw it away when it's done the job it was made for. It can be used in an experimental set-up at frequencies which diverge by up to 0.5%, even with another upper harmonic. It may also be used in a filter or demodulator.

The pre-requisite is that all its characteristics are known. The article describes ways in which they can be measured. This then leads to some new types of application. I assume that the reader is informed about the basic characteristics of crystals, as they are explained in the catalogues of most reputable manufacturers. What follows below is what you won't find there.

1. INTRODUCTION

Everyone knows this effect. You introduce a crystal into a tried and tested circuit and find that the frequency generated does not correspond to the one printed on the crystal. This can be verified using simple frequency meters, which nowadays measure with greater precision than that with which crystals are manufactured.

So who is at fault, the user or the manufacturer? Probably both! The user, because he or she is using the crystal in a way which the manufacturer did not expect, and the manufacturer, because he or she has not printed enough data on the crystal.

There are certainly enough examples of this. There is a crystal in my DIY box which is simply marked "60.2 MHz". The manufacturer apparently wishes to remain anonymous, and perhaps with good

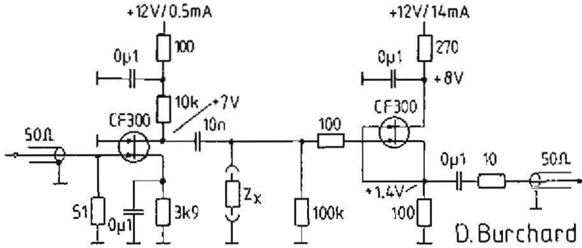


Fig.1:

Meaurement circuit for Impedances from 0 to app. 10kΩ; 50Ω Input and Output

reason. Another carries the description "6 MHz SWS". The abbreviation refers to a dealer who probably knows nothing else about the crystal. A third is referred to as "Valvo 18.000000". We can easily interpret this as 18 MHz (which is correct), with a comparison precision of 5 decimal places after the last digit written down (which is incorrect). It would actually mean that the tolerance was only 0.028 ppm. No manufacturer, however crafty, could promise that!

One example of a crystal marked with sufficient information is the following: TQ790516:216.631.25 range. First comes the code for the manufacturer, Telequarz (TQ). We then find that this is a crystal in an HC-45U housing (7...) in the ninth upper harmonic (.9...) with a comparison

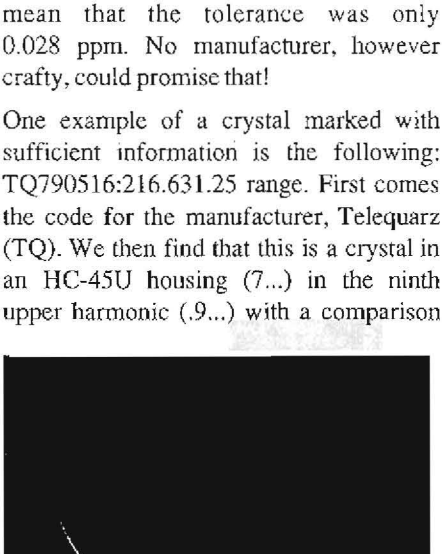


Fig.2: Impedance curve of a crystal Y: 5db/div; x: 25 MHz/div 0 ... 250 MHz

tolerance of ± 10 ppm (...05..) and TK of ± 7.5 ppm between -20 and +70°C (...16), which has series resonance at 216631.25 MHz. This is enough for the crystal to be used in accordance with the regulations. And there is hardly room for anything else in the small housing. If you want to use this crystal in a different way or to understand the cheap crystal with the inadequate description better, then you must do your own measuring.

2. CRYSTAL RESONANCES

Like any three-dimensional image, a crystal has more than one resonance. You can find any number of resonances, provided only that you make the frequency range of the investigation wide enough and select a sufficiently fine resolution. A suitable method is to measure the apparent resistance over the frequency range in question. Fig.1 shows a suitable measurement circuit, consisting of a power source and a source follower, which follows (1). It is wired into the signal path of a wobbler, network analyser or spectrum analyser with a tracking generator. For the TQ crystal referred to above, for example, you get an image like Fig.2. We are initially struck by a hyperbolic impedance

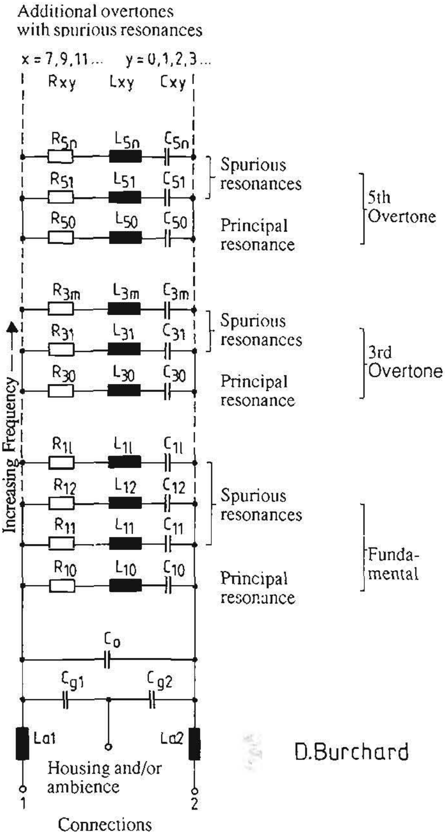


Fig.3: Complete equivalent circuit diagram of a Crystal

decrease, which in reality is a logarithmic hyperbola. It can be traced back to the capacity between the electrodes. The actual resonances of the crystal can be detected through small "swerves" in this hyperbola. This kind of thing is clearly visible at 24, 72 and 120 MHz, but can scarcely be made out at 170 and 217 MHz. The resolution is evidently not fine enough.

There are measuring instruments nowadays which can measure such curves in steps of 10 Hz. An expression of the frequency range of Fig.2 with a 0.1 mm. step width would be 2.5 km. long. So it is better to investigate each "jump" separately by reducing the sweep hub, once a general view has been obtained as per Fig.2.

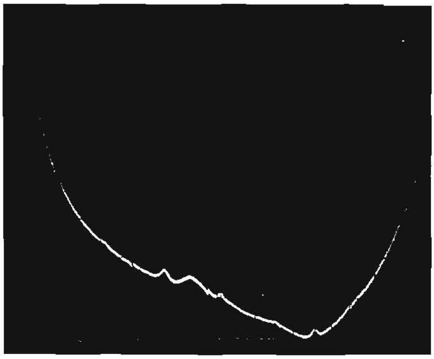


Fig.4: Appearance of a Zero Point in the Impedance Curve when long leads are used (scale division as Fig.2)

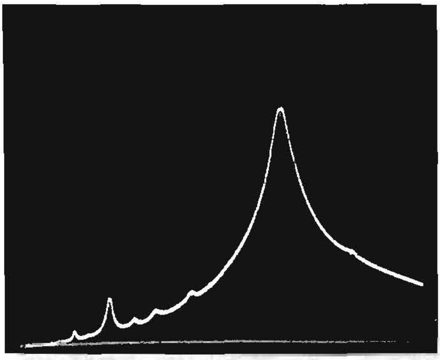


Fig.5: Compensation of electrode capacity generates a Parallel Resonance (here) atthe 7th Overtone (scale division as Fig.2)

There is thus a fundamental - this is the jump at the lowest frequency. Then follow the upper harmonics, each at an interval of about double the fundamental frequency. All of them have spurious emissions in their neighbourhood, as we shall see later. This gives the complete equivalent-circuit diagram of a crystal (Fig.3). Each resonance is represented by a circuit in series, R_{xy}, L_{xy}, C_{xy}. The capacity between the electrodes is referred to as C_0. Other capacities arise between the electrodes and the housing, which should usually not be neglected, and in addition the wires to the connections have a certain inductivity. The latter is obviously noticeable only in very long lines. For Fig.4, jumper wires 100 mm long were mounted as an extension at 10 mm. intervals, which created a series resonance with C_0 and C_g at 180 MHz. It can be concluded from this that L_{a1} and L_{a2} play no part in reasonable lengths of wire of under 10 mm., unless the frequency exceeds 300 MHz. At the moment, however, this is also the limit beyond which crystals are not available. It had, of course, already been reached 60 years ago with tourmaline crystals.

The representation of a series resonance is made considerably more difficult if there is a parallel low impedance of C_0. This is the case for all the higher upper harmonics. One way out is to compensate for C_0 by means of a parallel-wired coil. Its effect can be seen in Fig.5. The seventh overtone, for which compensation has been provided, is now clearly prominent, and the resolution can be markedly improved by reducing the sweep hub. The main resonance of the seventh overtone can be seen on the far left in Fig.6. All "satellites" to the right of it are spurious emissions. Now it is also clear that the resolution can not be increased any further using the broad-band wobbler. Its interference dispersion is several kHz, which is too wide. This means that steep parts of the curve are shown broken up and their more precise course or the minimum value (the series resonance) can not be made out.

3. MORE PRECISE REPRESENTATION OF RESONANCES

So a generator with a much narrower interference dispersion is needed; 100 Hz may be sufficient in many cases, 10 Hz always is. Such values can be found in signal generators which are provided for the measurement of high-quality receivers. This need not be the newest, most expensive model from one of the big manufacturers. What is required is the ability to modulate the frequency, with a lower limiting frequency as low as possible (1 Hz, 0 Hz are even better!), fine adjustability of the frequency and little drift, so that a meter can be connected up to determine the frequency. 10 Hz drift over a few seconds meets all requirements.

Old valve equipment from the surplus market with mechanical elements which are still operating smoothly can be modified relatively simply. The author has done this previously, using equipment from Rohde & Schwarz (SMAF type, made in 1954) and Eicke & Bemmerer (type SGU 702, made in 1965) with good results. A meter output was provided, electronic fine tuning was incorporated, and the double capacitors in the FM modulation path were

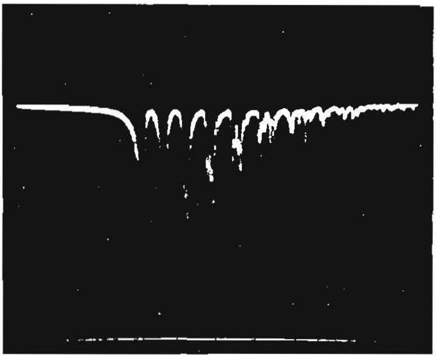
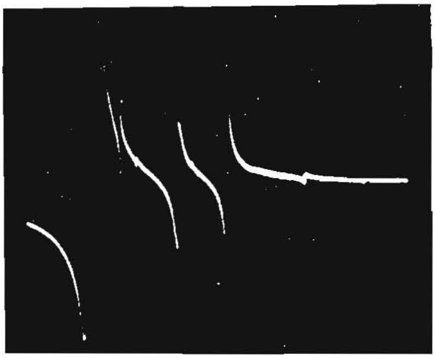


Fig.6: Main Waves and Spurious Emissions at 7th Overtone after Compensation



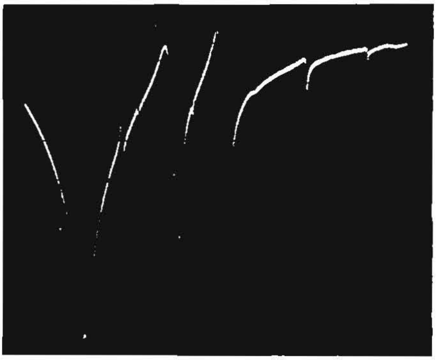


Fig.7: Principal and Spurious Resonances at Fundamental; (a)without compensation; (b) with compensation Y: 5dB/div; X: 30 kHz/div

considerably enlarged. However, the illustrations show the results obtained using a home-made signal generator, which can be modulated from DC and has an interference dispersion of about 50 Hz.

With such a generator and still rather a wide dispersion, we can obtain a good general view of the main resonances and their spurious emissions, shown here in Figs. 7 to 9 for the fundamental and the fifth and ninth overtones of the specimen TQ crystal. The spurious emission damping can be read off directly. Anyone not having a calibrated Y axis must change the generator voltage until first the principal wave and then the spurious emission coincide with the same screen line, and finally convert the results into dB.

The images shown in section a are obtained using uncompensated measurement. Here the parallel resonance can be seen just above the series resonance. It occurs at points where the imaginary part of the crystal series resonance impedance curve, in terms of quantity, is equal to the impedance of all C_0, C_g and external "load" capacities parallel to the crystal. The frequency is thus not determined through the crystal alone, but can be stretched rather far through external wiring. We do not need to determine it, because it can always be calculated from the equivalent data and the wiring. The images shown in section a also illustrate how measuring the series resonances without compensation is possible only for fundamentals and overtones of a low order. It is therefore a good habit always to measure using compensation. The criterion for correct compensation is a symmetrical shape for the series resonance curve in the vicinity of the resonance point, as can also be seen from the images in section b.

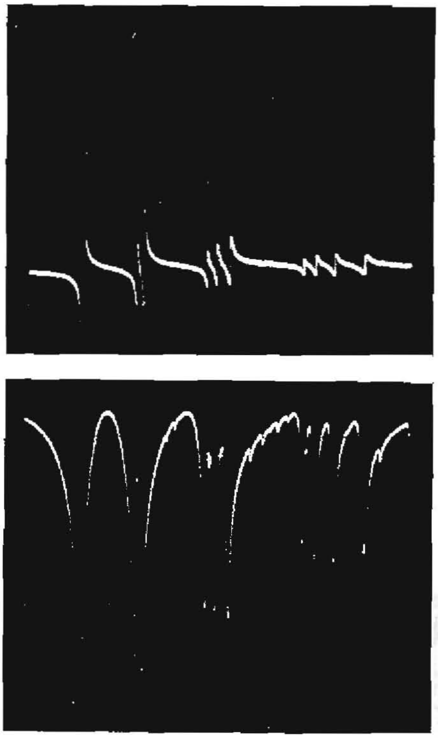


Fig.8: Principal and Spurious Resonances at 5th Overtone; (a) without compensation; (b) with compensation Y: 5dB/div; X: 20 kHz/div

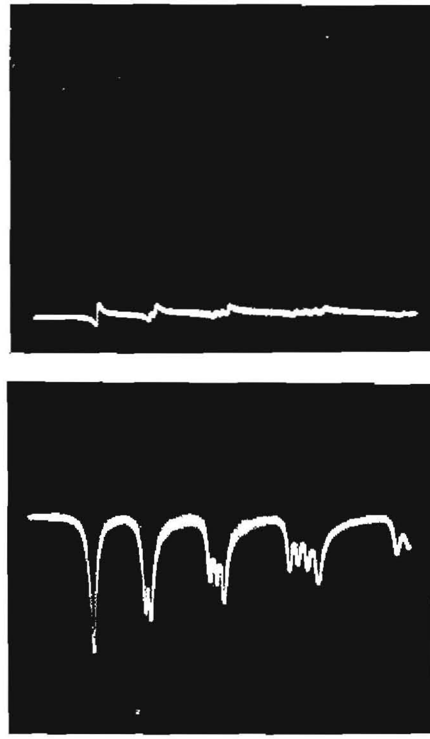


Fig.9: Principal and Spurious Resonances at 9th Overtone; (a) without compensation; (b) with compensation Y: 5dB/div; X: 20 kHz/div

The spurious emissions from the overtones occur at an increasing distance from the principal wave in groups of 2, 3, 4, etc. I have never yet found an explanation for this undoubtedly interesting behaviour anywhere.

4. MEASURING THE EQUIVALENT DATA

The electrode capacity, C_o, and the housing capacities, C_g, can be measured using any capacity measuring equipment with a sufficiently high resolution. The author uses a Marconi TF2700 universal Wheatstone bridge, which requires a measuring frequency of 1 kHz. This low frequency ensures that a crystal resonance is never measured when excited, and thus incorrectly. For capacity measurement, a crystal has three connections: the electrode leads and the housing. So three measurements must be carried out, each involving two connections linked to one another, so $C_x = C_{g1} + C_{g2}$, $C_y = C_o + C_{g1}$ and $C_z = C_o + C_{g2}$ can subsequently be determined. A simple calculation then gives:

$$C_o = ½(-C_x + C_y - C_z) \quad (1a)$$

$$C_{g1} = ½(C_x + C_y - C_z) \quad (1b)$$

$$C_{g2} = ½(C_x - C_y + C_z) \quad (1c)$$

Because of the symmetrical structure, one would expect that $C_g = C_{g2}$, and in the majority of cases this is correct. However, discrepancies do occur if, for example, the crystal is mounted slightly askew. The following capacities were measured for the test crystal: C_o = 4.1 pF; C_{g1} = 0.6 pF; C_{g2} = 0.7 pF.

With regard to the spurious emissions, the first point of interest is the spurious emissions interval, which has already been determined. Then we may perhaps wish to note the frequency intervals to the principal resonances. They can be read off directly, e.g. from Figs. 7 to 9. As a rule, we shall not want to know the L_{xy} and C_{xy} values of the spurious emissions. If we do, we proceed in exactly the same manner as described below for the principal resonances.

The desired resonance is brought into the centre of the screen by means of transmitter tuning, and the dispersion is reduced until an image corresponding to Fig.10 or Fig.11 is obtained. The symmetry indicates that the compensation has been set correctly. A suitable X frequency must be selected, so that the forward and reverse cycles coincide (splitting, as later in Fig.21, is only just permissible). The meter connected to the transmitter will eliminate the frequency variation by averaging if the gate time is long enough (e.g. 1 s.). This can easily be subsequently verified in that sequential meter displays vary by only a few Hz. The horizontal line in Fig's.10 and 11 corresponds to an impedance of 100Ω. If a 100Ω resistance is wired up instead of Z_x and another channel of the oscillograph is made to coincide with this line and is left there unaltered, we have this permanent reference line.

With the desired resonance exactly in the centre of the screen, the value is now read off from the meter. This gives the series resonance, f_{xo}. The +3 dB bandwidth has to be determined next. It is read off at the calibrated axes, or the transmitter voltage is reduced by 3dB and displaces the tuning in such a way that now the curve goes through the point on the screen where the

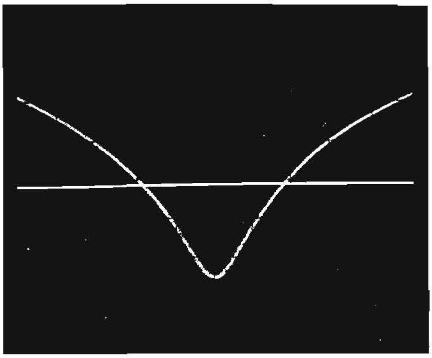


Fig.10: Image during measurement of Fundamental
Y: 5dB/div; X: 20 kHz/div

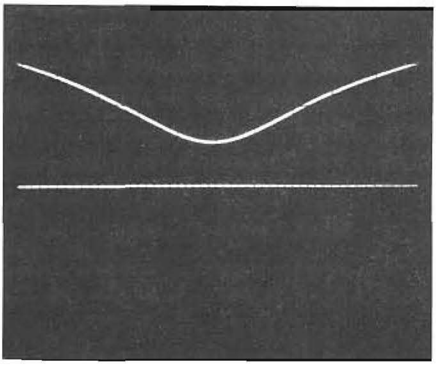


Fig.11: Image during measurement of 9th Overtone
Y: 5dB/div superimposed 100Ω line; X: 1 kHz/div

minimum value was previously. This gives two frequency values to be read off, their difference being the band width, fxo. The third value required is the series resonance resistance. By comparison with the 100Ω line and the calibrated Y axis, we obtain the factor in dB: -12 dB as regards Fig.10 or +6dB in Fig.11. The resonance resistance, R10, is thus 25Ω, while R90 is obtained as 200Ω. If there is no calibrated Y axis, the simplest method is to replace the crystal by a trimming potentiometer, set it in such a way that the line arising coincides with the valley point of the previous resonance curve, and then measure. Varying the transmitter voltage until the valley point coincides with the reference line also gives a result. We can then carry out the calculation:

$$Q_{xo} = f_{xo} / f_{xo} \qquad (2)$$

$$Lxo = \frac{Qxo \cdot Rxo}{2 \cdot \pi \cdot f_{xo}} \qquad (3)$$

$$C_{xo} = \frac{1}{2 \cdot \pi \cdot f_{xo} \, . \, Q_{xo} \, . \, R_{xo}} \qquad (4)$$

The value measured for the sample crystal, together with the equivalent data calculated from it using equations (2)...(4), is shown in Table 1. By comparing the data with those from the manufacturer (2), we can recognise that they are all within the given range. The obvious conclusion is that there is no basic variation in the manufacture of basic crystals and overtone crystals, at least not with this structure. And there is thus nothing to prevent the crystal's being used at another resonance.

A comparison of the f_{xo} frequencies measured discloses that the overtones are not distributed harmonically - a fact mentioned in every textbook. But what is not mentioned is that they are all distributed harmonically in relation to a fictitious keytone, which here is 24.070 MHz. If this is taken into account, the overtones can be calculated with an error amounting to only a few tens of ppm. There is an obvious explanation. The electrodes charge the crystal and lower its frequency (here by 1,255 ppm). This effect is used for the fine tuning of the frequency, in that metal is deposited until the rated frequency is reached. This influence is at its greatest for the fundamental. The batches of metal are located at the maximum amplitudes. The overtones can not be influenced so easily, because several maximum amplitudes lie within the crystal's interior. The fictitious fundamental is calculated by halving the frequency interval for two adjacent overtones.

		Fundamental x = 1	3. Overtone x = 3	5. Overtone x = 5	7. Overtone x = 7	9. Overtone x = 9	
Measured value	f_{xo}	24,040.5	72,211.6	120,356.4	168,493.4	216,629.6	MHz
	Δf_{xo}	1,40	2,60	2,80	4,20	4,20	kHz
	R_{xo}	25	50	65	130	200	Ω
Calculated value	Q_{xo}	17000	28000	43000	40000	52000	–
	L_{xo}	2,8	3,1	3,7	4,9	7,6	mH
	C_{xo}	15,6	1,6	0,47	0,18	0,07	fF

Table 1

5. MEASUREMENT OF TEMPERATURE BEHAVIOUR

The crystal oscillator has such a small thermal capacity that it very rapidly adopts the temperature of the housing due to radiation. If the housing in the measurement circuit is treated with a cooling spray or with hot air, the resonance can be seen to drift, and can also be measured by trimming the frequency at the centre of the screen. The cooling spray generates snow on the housing. If the snow melts, the temperature is pretty close to 0°C. A less well-defined temperature point lies at 50°C, ''measured'' by testing with a finger after heating. If the temperature is still higher, then you will have to move your finger away at once. At 50°C a brief contact is already possible. Anyone wanting more precise information will have to put a thermometer over the crystal or use a digital thermometer.

The two symmetrical temperature test points located at 25°C make it possible to determine whether the crystal was manufactured for a wide temperature range (-20 to +70°C) or a narrow one (0 to +50°C), or for a specific temperature (thermostat operation 60...90°C). The crystal catalogues usually give temperature behaviour curves which allow the three cases to be distinguished from one another. Then again, there are examples for which the frequency decreases below and above 25°C. This involves, not a thermostat crystal for 25°C, but a BT cut. Many cheap manufacturers prefer this cut for frequencies of 15...30 MHz, because the crystal is thicker than for an AT cut, and so easier to produce. The disadvantage is a considerably higher TK, which restricts use to the less critical cases, e.g. micro-processors.

6. CHANGING THE RESONANCE FREQUENCY

This procedure is referred to as ''stretching the crystal'', and is considered as being difficult and possible only to a small extent. This is definitely wrong. Later, we shall make the acquaintance of circuits which allow considerable ''stretching''. It is even true that the natural limit for stretching is at fairly high frequencies, because spurious resonances always arise there. The stretching procedure would have a point of discontinuity!

Stretching is always required if a circuit has to supply a more precise frequency than is given by the manufacturer's tolerance for the crystal. Otherwise, it would not be possible to align a frequency standard to 0.01 ppm ($1 \cdot 10^{-8}$) using a crystal with an accuracy of adjustment of 2 ppm.

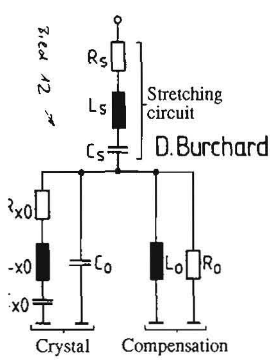


Fig.12: Universal stretching circuit for Series Resonance Crystals

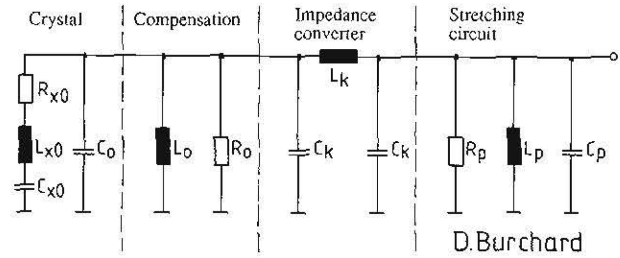


Fig.13: Universal Stretching Circuit for a Parallel Resonance, conversion of Series Resonance into Parallel Resonance

Fig.12 shows a general "stretching circuit" for series resonance. It consists of a second series resonance circuit (R_s, L_s, C_s), which is wired up in series with the crystal. For symmetrical stretching characteristics, the crystal must be compensated for. Initially, the stretching circuit must have the same resonance frequency as the crystal. Detuning them leads, in accordance with equation (5) below, to the displacement of the total resonance by Δf_z. In practise, the possible Δf_z changes are still very small, as against the crystal resonance, f_{x0}; there is a simple interrelationship:

$$\Delta f_z = f_{zo} \cdot \frac{C_{xo}}{2} \cdot \left(\frac{1}{C_s} - 4 \cdot \pi^2 \cdot f^2_{xo} \cdot L_s\right) \quad (5)$$

Thus, if Ls is missing from the circuit, only a stretching capacitor is wired in series, and the equation is simplified to:

$$\Delta f_{zc} = f_{x0} \cdot \frac{C_{xo}}{2 \cdot C_s} \quad (5a)$$

Frequencies can then only be increased.

If, on the other hand, the circuit contains only a stretching coil, then the equation which applies is:

$$f_{zL} = -2 \cdot \pi^2 \cdot f^3_{x0} \cdot C_{x0} \cdot L_s \quad (5b)$$

and frequencies can only be reduced.

The network theory states that a network made up of n reactive components has n - 1 resonance points. The crystal alone has three reactive components (C_{x0}, L_{x0}, $C_0 + C_g$), and thus two resonances; namely, the series resonance, which is always used here, and the parallel resonance, which is just above it, without compensation. If L_0 is introduced, a further resonance arises. It can be seen from Fig.7 that the original (natural) parallel resonance is displaced to considerably higher frequencies, whilst a new series resonance arises below the existing value. Two further series resonances arise above and below the existing parallel resonance points. Without further measures, their resonance resistance can be below $R_{x0} + R_s$, and an oscillator can be excited at one of these resonances. This can be remedied by reducing R_0. This resistance should initially represent the losses from L_0. If it is artificially increased, the parasitic series resonances are dampened.

A universal stretching circuit for the natural parallel resonance of a crystal does not exist, because this parallel resonance comes into being only through the interaction of further components and is not a characteristic of the crystal. However, by converting the impedance it is possible to convert a series resonance into a parallel resonance. The best-known impedance converter would probably be a λ/4 long piece of coax cable. But there are also solutions using discrete structural elements. Fig.13 shows one such. Its conversion equation is:

$$Z_{in} \cdot Z_{out} = \frac{L_k}{C_k} \quad \text{when } L_k \cdot C_k = \frac{1}{\omega^2_{x0}} \qquad (6)$$

The impedance curve of the compensated crystal is thus inverted. A parallel-wired parallel stretching circuit acts in a similar way to the series circuit in Fig.12. The circuit is relatively complicated, even if some components are combined (C_0 and C_k, or C_k and C_p). Although this is a solution which satisfies the purists, the circuit is nevertheless very seldom used. So the calculation formulae for it will be omitted.

It is frequently sufficient to allow the current in a series circuit to flow through a resistance so as to obtain something like a parallel resonance. This frequently happens with uncompensated crystals which are under capacity load. We have a further example in Fig.18.

7. APPLICATIONS

There are a great many circuits for crystal oscillators in the literature. There is no space here to describe or evaluate them. I shall therefore limit myself to three specimen circuits, which are particularly good illustrations of the above remarks. All three use compensated crystals. To complete the picture, it should be explained that, in addition to a parallel coil, such compensation can also be provided by a capacity equal to C_0, which is powered by a counter-phase voltage. You need amplifiers with a counter-phase output or with a differential input, and solutions using differential transformers and bridge circuits are also conceivable. In any case, increasing attention must be paid to this problem as the frequency rises.

The next question is whether the crystal housing should be connected somewhere. The capacities between it and the electrodes are not negligible. Earthing is a good solution when one electrode is already earthed anyway or when the two are connected through earthed capacitors. If you decide not to connect up the housing, any foreseeable earth contact must also be avoided. An intermittent contact between the crystal housing and the earth surface of the printed circuit board can cause unattractive effects.

The circuit in Fig.14a is a crystal-stabilised phase circuit for a quadrature demodulator. This type of FM demodulation appeared on the scene with integrated circuits and requires a parallel resonance for two reasons. The circuit is powered from a

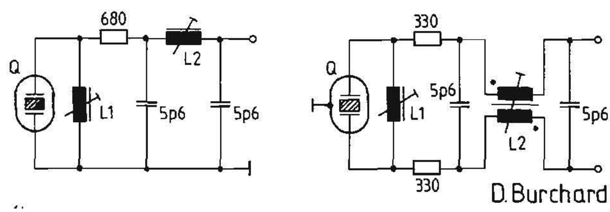


Fig.14: Crystal-Stabilised Parallel Resonance Circuit for use in an FM Quadrature Demodulator
(a) Asymmetrical, NE604
(b) Symmetrical, TDA1576

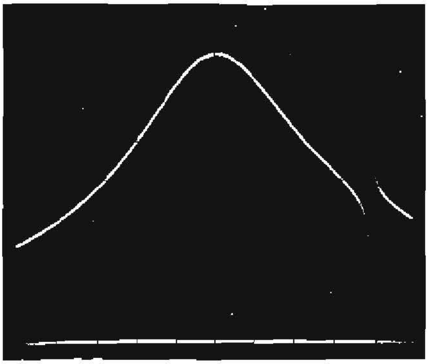


Fig.15: **Fundamental Crystal Parallel Resonance generated in a circuit as per Fig.14. Y: linear, 0 line superimposed X: 10 kHz/div, average frequency 24040.4 MHz**

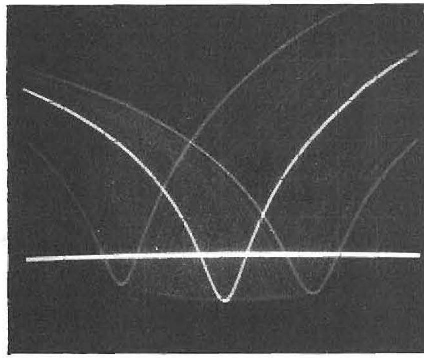


Fig.17: **Operation of Fig.16 circuit Y: 5dB/div, 100Ω line superimposed X: 5 kHz/div, average frequency 24053 MHz**

source which can not be subjected to whatever load you wish, and the input of the phase monitor requires a sufficiently high voltage. A circuit in series resonance would provide the same phase displacement in itself, but is ruled out for the reasons specified above. Naturally, there are many types of circuit which are also in operation somewhere. The circuit shown here is a "precise" solution to the problem.

The crystal compensated with L1 is taken to a sufficiently high band width, using a fixed resistance. This series resonance is

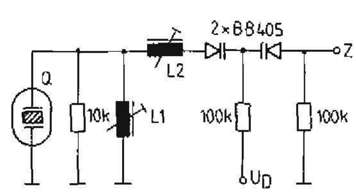


Fig.16: **Circuit of a Crystal Series Resonance which can be electronically stretched**

converted into a parallel resonance at 25 kHz using an impedance converter. There is thus a hump interval of 25 kHz in the quadrature demodulator and a useful linear band width of app. 15 kHz. The resulting parallel resonance can be measured using a circuit as per Fig.1. Fig.15 shows the measurement curve. An (inverted) secondary resonance can also be seen here, which lies outside the operating range. Many integrated quadrature demodulators need an asymmetrical structure. The part circuit shown in section a of Fig.14 is to be used in this case. Others, such as the well-known TDA 1576, require a symmetrical circuit, as represented by the part circuit in section b. The average frequency coincides with the series resonance frequency of the crystal. In practise, the two coils are tuned in such a way that the average frequency can be changed to a slight degree using L2 (but exceeding the tolerance of the crystal), whilst L1 ensures the linearity alignment.

The circuit in Fig.16 shows a resonator which, although crystal-stabilised, can still

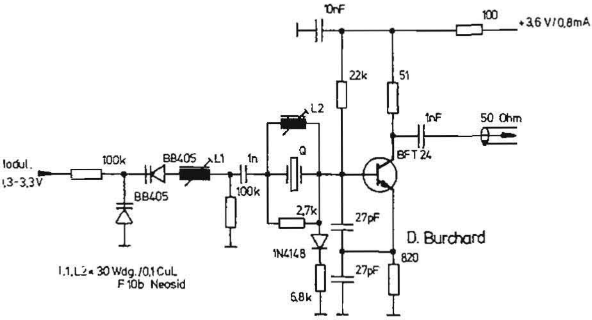


Fig.18: FM Oscillator circuit, Frequency 24053 MHz, Dispersion ± 12.5 kHz

have its frequency varied to the extent of ± 500 ppm. It is used, for instance, in an oscillator which has to be "connected" to a considerably more precise standard frequency, or for fractional detuning, in a system which is otherwise digitally operated and does not allow for sufficiently fine frequency steps.

The inductivity, L1, compensates the crystal, the 10kΩ resistance dampens unwanted serial resonances, L2 and the capacitance diodes form the stretching circuit. The impedance curve can be measured at output Z as per Fig.17. For

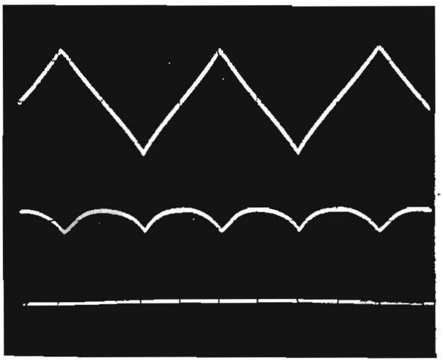


Fig.19: Modulation quality of Fig.18 circuit
Y1: 10 kHz Dipersion/div
Y2: 50% AM/div, zero line of rectifier superimposed
X: 0.5ms/div

this option, a DC voltage was fed into U_D, and a square-wave AC voltage was then superimposed on it for maximum frequency variation. During the exposure time, the square-wave voltage was slowly turned down to zero. The almost symmetrical detuning and the very high constancy of the resonance resistance can be recognised. The latter is naturally higher than that for the crystal alone, as is shown by a comparison with the 100Ω line, because the loss resistance of the stretching circuit is added. The resonance curve itself is markedly asymmetrical. The reason for this is that the capacity curve of the C-diodes plotted against the voltage does not match equation (5). It is not hyperbolic, but is more in accordance with the equation $C = C_0 (U + U_D)^{-m}$; U_D is about 0.6V and m is between 0.3 and 0.7, depending on the manufacturing process. The non-linearity of the frequency variation, restricted in this way, can be almost compensated for by slightly detuning L1. Of course, the C_0 compensation of the crystal is no longer correct. So in practise the linearity is set using L1 and the average frequency desired is set through L2.

The third circuit, Fig.18, represents the oscillator circuit of an FM transmitter, which was developed for scientific purposes. After quadrupling, it supplies wide-

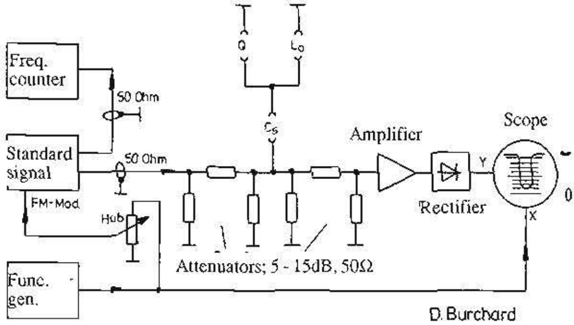


Fig.20: Simple Crystal measurement circuit

and FM over the standard radio range. Here the series resonance of the crystal is converted into a parallel resonance using an input capacitance, making the average frequency 12.5 kHz higher. In other respects, the circuit is largely the same as the one in Fig.16. However, the sequence of the components in the oscillation circuit is slightly altered, and a modulation input and a 50Ω output are created using a transistor.

The question of earthing the crystal housing arises again here. The author decided in favour of insulation and for safety reasons covered the crystal with a heat-shrinkable tube before assembly. Fig.19 testifies to the good quality of the modulation. According to the oscillator, the synchronous amplitude modulation fraction is only 15%, and largely disappears in the two subsequent doubler stages. The frequency modulation is noticeably linear, although the modulating voltage uses almost the entire operating voltage range. Here again, the precise frequency is set using L1 and the linearity is set using L2.

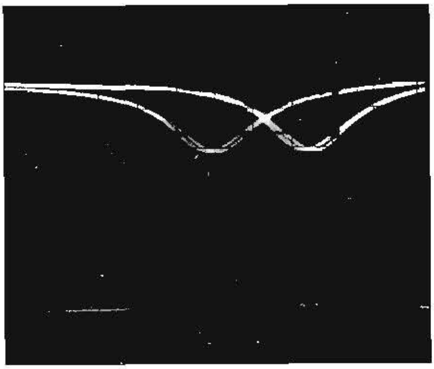


Fig.21: Screen Photo: Crystal Measurement using Fig.20 circuit and hand-written special scale
Y: 0.... ∝Ω, 0Ω superimposed
X: 2 kHz/div

8. IT IS EVEN SIMPLER TO MEASURE THE CRYSTAL

It should be remembered that a crystal with a series capacity, C_s, has the same series resonance as the same crystal in parallel resonance with a parallel capacity where $C_p = C_s$. So if the parallel resonance can also be measured as a series resonance, there is no longer any need for a measurement circuit as per Fig.1, suitable for measurements even in the kΩ range. Thus a considerably simplified circuit, as per Fig.20, is conceivable.

Here the crystal is mounted between two attenuators, which are there to ensure that the generator and amplifier see constant system resistances. Where the crystal is mounted, the impedance is 25Ω. The amplifier can consist of a wideband gain block. It is intended to compensate for the losses in the attenuators and simultaneously to add so much amplification that a simple rectifier can supply enough output voltage for an oscilloscope. The type of rectification and the curvature of the characteristic lines are of no importance. They are "calibrated into" the display on the oscilloscope. Only the same adjustment range (V/div.) must always be selected there.

The "calibration" is done with a felt-tip pen on an overhead film in front of the screen. The zero line is set on the oscilloscope without any generator voltage. The generator voltage is then turned up until the amplifier is well modulated but still in the linear range. Both the lines visible on the screen are marked, one as 0Ω, the other as ∝Ω. If we now switch in resistances of 10, 20 and 50 one after the other up to 500Ω, we obtain a calibrated Y scale. Intermediate values are estimated later or determined by means of substitution.

The procedure for measuring the resonance of any crystal is as follows:

- → Set the approximate frequency
- → Set the zero line without generator voltage
- → Increase generator voltage until the line is reached
- → Insert crystal and a short-circuit instead of C_S
- → Trim generator frequency to principal resonance
- → Insert compensation coil and adjust to best symmetry
- → Keep the resonance in the middle of the screen and read off the series resonance frequency on the meter
- → Read off the resonance resistance
- → Insert known stretching capacitor, C_S
- → Bring resonance to middle of screen again, read off the frequency and calculate the stretching frequency

The process is illustrated in Fig.21. The "graduations" in Ohms can be vaguely recognised in front of the screen, and a 0Ω line is also superimposed. The line is reached outside the resonance to be measured. During the exposure, the previously short-circuited 5pF stretching capacitor is switched in, and the resonance jumps from the middle of the screen to a value app. 5 kHz higher. The resonance resistance of app. 65Ω remains constant while this is happening, because of the compensation. This picture was taken with the specimen crystal on the fifth overtone.

Two defects in the image in Fig.21 require some explanation. An additional resonance can be seen very faintly at +7 kHz. This arises while the stretching capacitor is being reversed due to the switching capacity. Then a pronounced split can be seen in the curve between the scanning and the return passes. This is due to the increased recording speed. In order to ensure a certain basic brightness for the screen, so that the scale could be seen on the film, the X voltage was selected to be ten times as high as was necessary. So for 90% of the time the electron radiation falls on the glass wall of the valves and generates secondary electrons there. These reach the screen and generate a diffuse brightening.

The equivalent data are calculated using somewhat re-modelled versions of the formulae already given above:

$$C_{x0} = \frac{2 \cdot C_{zC} \cdot \Delta f_{zC}}{f_{x0}} \qquad (7)$$

$$L_{x0} = \frac{1}{8 \cdot \pi^2 \cdot f_{x0} \cdot C_{zC} \cdot \Delta f_{zC}} \qquad (8)$$

The spurious emissions interval is obtained by measuring the resonance resistances of the spurious emissions and converting them into dB:

$$NWA/dB = 20 \log \frac{R_{xy}}{R_{x0}} \qquad (9)$$

It should be mentioned here that this type of crystal measurement is not according to standard (DIN 45105), but against this is considerably simpler. No special measuring head is required, no compensation branch and no phase meter. I see a further advantage in the fact that the capacities C_{g1} and C_{g2} can be taken into account in unambiguous manner by the single-ended crystal. If the housing is earthed, C_{g1} is parallel to C_0 and C_{g2} is short-circuited.

9. LITERATURE

(1) Burchard, D. (1992): MES-FETishism II
V F Communications 4/1992, pp.223-240

(2) Telequartz Catalogue technical Introduction ff, Neckarbischofsheim

Gunther Hofmann, DK 2 TX

Programming the DSP Computer

This short article was inspired by the recognition that the programming language for the SDP computer needs a lot of getting used to. Moreover, due to the decided scarcity of literature on the DSP program, a lot of effort has to be put into understanding the program well enough to put oneself in a position to be able to make amendments to it, let alone write such programs oneself.

```
< ------------- Check and Change Memory CCMEM ------------ >
<                                                          >
< The program CCMEM is used to list and change the contents>
< of data areas of the DSP computer.                       >
<                                                          >
< - The program is started by the command R CCMEM.         >
< * First CCMEM asks for the start address.                >
< - Type the start address (6 hexa characters max) followed>
<   by a 'CR'.                                             >
<   If you have typed an odd address, CCMEM takes the first>
<   even address below the input one.                      >
< * Now you are asked to type the end address.             >
< - Type the last address of the memory area to be listed  >
<   (6 hexa characters max.) followed by a 'CR'.           >
<   If you have typed an odd address, CCMEM takes the first>
<   even address below the input one.                      >
<   If the 'End address' is less than the 'Start address'  >
<   an error message will be printed and CCMEM will        >
<   terminate.                                             >
< * After the second input the desired memory area will be >
<   listed on the screen from the started address up to the>
<   end address.                                           >
< * When the last address is output, CCMEM is waiting for  >
<   further inputs:                                        >
< - '-'     :CCMEM shows the contents of the current address>
<            minus 2 (which is the previous address).      >
< - 'blank':CCMEM shows the contents of the next even      >
<            address (which is the current address plus 2). >
< - 'CR'    :CCMEM terminates.                             >
< - '4 character hexa number': CCMEM replaces the contents >
<            of the current address by the new hexa number. >
<                                                          >
< Common behaviour on memory content's inputs:             >
< - CCMEM ignores non hexa characters in memory content's  >
<   inputs.                                                >
< - If less than 4 characters are input, the number is     >
<   interpreted 'right hand justified' e.g. 200 means  0200>
< - If more than 4 characters are input, only the last 4   >
<   ones are taken.                                        >
< This behaviour can be used to correct erroneous inputs   >
< just by repeating the last 4 characters.                 >
<                                                          >
< Example:                                                 >
< Note: {CR}  means 'carriage return'                      >
<       {BLK} means 'blank'                                >
<       ***   means comment                                >
< R CCMEM{CR}                                              >
<·Start addr: 200000{CR}          *** Input of start addr  >
< End   addr: 200004{CR}          *** Input of end addr    >
< 200000 AA46                     *** Output of listing    >
< 200002 4654                                              >
< 200004 2020 34141{BLK}          *** Input of new contents>
< 200006 2020-                    *** List previous address>
< 200004 4141{CR}                 *** List new contents    >
< .                               *** CCMEM has terminated >
< -------------------------------------------------------- >
```

Fig.1: CCMEM Utilityprogram

```
<***** YT3MV - RTTY AFSK Asynchronous RX/TX - 03/12/1990 ****>
<********** Additional Comments by DK2TX 17/04/92 ***********>

< -------------------- Common Constants ------------------- >

32768*65536=npr            <Value to force integer calculation >
-1$d=pi                    <For filter coefficients computing  >
9600=frs                   <Sample frequency                   >

< ----- Initialisation of uPD71054 for a frs = 9.6 kHz ----- >

&<13FC 0016 0002 4007>   <move.b #$16,$24007; 71054 control  >
&<13FC 0010 0002 4001>   <move.b #$10,$24001; A/D Sampling   >
&<13FC 0096 0002 4007>   <move.b #$96,$24007; 71054 control  >
&<13FC 0010 0002 4005>   <move.b #$10,$24005; D/A Sampling   >

< -------- Values to generate the CODEC lookup table ------- >

<The coding of the CODEC Data is as follows:                 >
<inv: inverted                                               >

< +------+------+------+------+------+------+------+------+  >
< |           Mantissa          |      Exponent      | Sign |  >
< | inv |        | inv |        | inv |      | inv   |      |  >
< | LSB   ------------}  MSB    | LSB ---------} MSB |      |  >
< +------+------+------+------+------+------+------+------+  >

$:cho(7)                   <CODEC chord values               >
0=cho(5)  16.5=cho(1)  33=cho(7)   66=cho(3)
132=cho(4) 264=cho(0) 528=cho(6) 1056=cho(2)

$:ste(7)                   <CODEC steps                      >
 1=ste(5)  1=ste(1) 2=ste(7) 4=ste(3) 8=ste(4) 16=ste(0)
32=ste(6) 64=ste(2)

$:nst(15)                  <CODEC step numbers               >
 0=nst(10) 1=nst(2)  2=nst(14)  3=nst(6)  4=nst(8)   5=nst(0)
 6=nst(12) 7=nst(4)  8=nst(11)  9=nst(3) 10=nst(15) 11=nst(7)
12=nst(9) 13=nst(1) 14=nst(13) 15=nst(5)


< ---------- BAUDOT to ASCII conversion table ------------- >

$:t(1,31)                  <Lower case/upper case            >

 0=t(0,0)   0=t(1,0)  69=t(0,1)  51=t(1,1)     <NIL,NIL, E, 3>
10=t(0,2)  10=t(1,2)  65=t(0,3)  45=t(1,3)     < LF, LF, A, ->
32=t(0,4)  32=t(1,4)  83=t(0,5)  42=t(1,5)     <   ,   , S, *>
73=t(0,6)  56=t(1,6)  85=t(0,7)  55=t(1,7)     <  I,  8, U, 7>
13=t(0,8)  13=t(1,8)  68=t(0,9)  36=t(1,9)     < CR, CR, D, $>
82=t(0,10) 52=t(1,10) 74=t(0,11) 39=t(1,11)    <  R,  4, J, '>
78=t(0,12) 44=t(1,12) 70=t(0,13) 33=t(1,13)    <  N,  ,, F, !>
67=t(0,14) 58=t(1,14) 75=t(0,15) 40=t(1,15)    <  C,  :, K, (>
84=t(0,16) 53=t(1,16) 90=t(0,17) 34=t(1,17)    <  T,  5, Z, ">
76=t(0,18) 41=t(1,18) 87=t(0,19) 50=t(1,19)    <  L,  ), W, 2>
72=t(0,20) 35=t(1,20) 89=t(0,21) 54=t(1,21)    <  H,  #, Y, 6>
80=t(0,22) 48=t(1,22) 81=t(0,23) 49=t(1,23)    <  P,  0, Q, 1>
79=t(0,24) 57=t(1,24) 66=t(0,25) 63=t(1,25)    <  =,  9, B, ?>
71=t(0,26) 38=t(1,26)  2=t(0,27)  2=t(1,27)    <  G,&,STX,STX>
77=t(0,28) 46=t(1,28) 88=t(0,29) 47=t(1,29)    <  M,  . ,X, />
86=t(0,30) 59=t(1,30)  1=t(0,31)  1=t(1,31)    <  V,;,SOH,SOH>

< --------- Stored options table & file name table --------- >
```

Fig.2a: Extracts from RTTY.src program with comments

The CCMEM program makes it possible to list memory areas by entering the start and end addresses, and to amend them simply (Fig.1). This has proved to be of great practical benefit, for example, in looking at the field surrounding a problem triggering an error message and correcting errors rapidly without amending the source language.

Those wishing to write programs for the DSP computer themselves and not all that familiar with the syntax will perhaps start with the analysis of existing programs. However, there is very little literature dealing with existing programs. This is why I have provided the RTTY.SRC program from YT3MV with more extensive comments (Fig.2). With these comments, the structure and format of the program can be reconstructed.

```
#443                   <    Command 20 and invalid numbers >
nva-20.5$j+400         <==} Invalid option number          >
$n$b$i=sfor            <    Set new number of columns      >
$j400                  <==} Back to main menue             >

#410
< =========== Subroutine for file name output =========== >

0=nch                  <Number of first char. (loop cntr.) >
#411                   <Loop for prefix output             >
f(nva,nch)$w           <Output of one letter of the prefix >
nch+1=nch              <Increment character number         >
nch-11.5$j-411         <==} Handle next character          >

$<.>                   <'.' between prefix and suffix      >

#412                   <Loop for suffix output             >
f(nva,nch)$w           <Output of one letter of the suffix >
nch+1=nch              <Increment character number         >
nch-14.5$j-412         <==} Handle next character          >
$r                     <=} Return from subroutine 410      >

#450
< = Subroutine to find TX file and to get address & length = >

$:name(2)              <Buffer for file name with 24 char. >
name(0)&<200A D08C 7200>=name <Get address of name(0)      >
                       <move.l  A2,D0;  Var. offset addr.  >
                       <add.l   A4,D0,  Var. base address  >
                       <moveq   #0,D1;  Reset buffer index >
0=nch                  <Character number (loop counter)    >
#451                   <Loop for file name transfer        >
f(0,nch)+npr&<2E00>    <Read one character from 'f(0,nch)' >
                       <move.l  D0,D7;  Save the character >
name&<2040 1187 1000 5241>=name  <Shift it to buffer 'name' >
                       <move.l  D0,A0;  Get buffer address >
                       <move.b  D7,0(A0,D1.W); Store char. >
                       <addq.w  #1,D1;  Increment index    >
nch+1=nch              <Increment char number (loop cntr)  >
nch-14.5$j-451         <==} Character transfer loop        >
name&<2040 4E49>=len   <move.l  D0,A0; Get name buffer addr>
                       <trap    #9;    Search file         >
&<2009>=add            <move.l  A1,D0; Save file address   >
$r                     <=} Return from subroutine 450      >

#500
< ============= RTTY Receiver/Transmitter ============= >

&<4E45>                <trap #5; Reset keyboard            >
-1=mfl                 <Initial state: Recording off       >
$:chb(999)             <Receiver character buffer          >
999.2=ch1=ch2          <Reset RX buffer pointers           >
$:txb(222)             <Transmitter character buffer       >
222.2=tx1=tx2          <Reset TX buffer pointers           >
-1=tuf                 <Switch off tuning indicator        >
$z100                  <=} Initialize RTTY receiver        >
-1=txf                 <Mark transmitter state as 'off'    >
-1=ffl                 <Mark file output as off            >
444=blp                <On/off time for cursor blinking    >

#505
< ------------- Clear screen & print header -------------- >
```

Fig.2b: Extracts from RTTY.src program with comments

LITERATURE

(1) Matjaz Vidmar, YT3MV; Digital Signal Processing Techniques for Radio Amateurs
Part 1: VHF Communications, 4/88
Part 2: VHF Communications, 1/89
Part 3: VHF Communications, 2/89
Part 4a: VHF Communications, 3/89
Part 4b: VHF Communications, 4/89

(2) Matjaz Vidmar, YT3MV; Rapid Fourier Transformation in Amateur Radio
Part 1: VHF Communications, 1/90
Part 2a: VHF Communications, 2/90
Part 2b: VHF Communications, 3/90

(3) Matjaz Vidmar, YT3MV; DSP-Computer Update No. 1
VHF Communications, 3/91

Dipl.-Phys. Nothart Rohde

EMC - and its Consequences

High-frequency equipment plays such a large part in modern electronics that it has to be dealt with in accordance with the well-known design principles of HF technology. Nowadays, standard flip-flops can already divide 100 MHz and the ensuing maximum noise zone is quite remarkable.

1. INTRODUCTION

In practise, the requirement for electromagnetic compatibility (EMC) means that however reasonably any piece of electrical or electronic equipment is designed, it is bound to need shielding and screening. Neither the internal signal flow within the equipment nor the external serviceability should be altered or impaired to achieve this - frequently a puzzle which simply has no solution.

Apart from the power electronics, the biggest problems are probably receiver-computer links and circuit power supplies. One or two of you have undoubtedly had experience of this yourselves.

Until now, most electronics specialists have had no understanding of the need for EMC measures. And we already have video applications operating at clock frequencies of anything up to 250 MHz!

Unfortunately, the manufacturers of these interfering transmitters, to which ring mixers can easily become connected, are scarcely in a position, or are scarcely willing, to understand the problems of perturbing radiation. On the other hand, a considerable increase in radio communication can be observed. In many cases, these radio connections actually have only become possible due to assistance from digital technology. Forward-thinking people are already talking of an age of communication. The astonishing progress in digital technology, for one thing, and the increasing requirement for undisturbed frequencies. for another, are generating a greatly expanding number of unsolved problems. Solving them will need competent specialists who understand both technologies and who can therefore act as consultants.

And with regard to the European unit, even standards of this nature lay down requirements which it should actually have long ago been possible to adhere to, or even to improve on. For experience teaches us that adherence to standards in no way guarantees the operability of equipment. Radio amateurs should be aware of that from the example of the beam strength of TV receivers.

The solution recommended by the relevant suppliers of EMC measuring equipment is the creation of a complete EMC measurement shed, which is big enough to take a car. On the other side are the purchasers from the companies affected, who lay great emphasis on being able to measure their products according to the standards, but also economically. Of course, no-one spares a thought for whether the technician in question can acquire the signal on site in accordance with the standard and make it available to the measuring equipment. It is naturally assumed that this technician has already worked through mountains of VDE guidelines, Ministerial decrees and EMC literature.

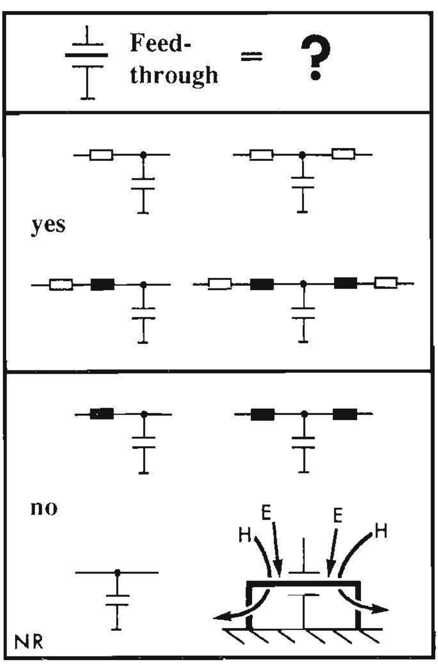


Fig.1: Diagrams for a feed-through capacitor

Even if we accept that EMC measurement gives us real facts and data to place on the table, that's still a long way from saying that the development specialists can draw the appropriate conclusions from them and can grasp suitable corrective measures.

Up until now, it has simply been awareness of the EMC problem which was lacking, and with it the readiness and the will to consider the EMC aspect of a development at all stages. The route taken up until now - testing a finished product and then making modifications to it, is simply unacceptable.

This can be demonstrated by numerous examples on the basis of which housings or which components a wholesaler has in stock, or on the fact that certain components relevant to EMC may admittedly only cost a few pence, but nevertheless are available only from specialist amateur radio equipment dealers.

EMC is the problem of all those who are involved with electronics or radio technology in any way. So radio amateurs should also be active in this area. In the end, we are talking about frequency ranges which will otherwise be made unusable.

EMC is a very complicated and far-ranging topic and it would be easy to write several books about it. There has been very little really helpful literature so far. I might mention Durcansky's book (1) and some work of my own on measuring equipment, aerials and general principles (2).

Let's stick to practical tips for experimental work. No doubt some points will already be familiar. Others are so simple and effective that they might at first be overlooked.

2. THE FEED-THROUGH CAPACITOR

It has already become established on safety grounds that high-frequency sub-assemblies should be screened and voltages supplied through feed-through capacitors (FTC's). It is hoped that unwanted irradiation and transmissions can thus be suppressed.

What is actually gained by doing this?

In principle, a feed-through capacitor is part of a low pass, which it is advantageous to supplement by resistances and inductances, whether to the half-section or the T-network. Various options are shown in Fig.1 using different equivalent circuit diagrams. Without these additions, the mode of operation is poorly defined. The simplest equipment to use consists of resistances which lead to limiting frequencies in the low-frequency range and can easily be de-coupled. Chokes should be provided on power supply lines due to the voltage drop. It should be borne in mind that the low pass is not complete without an additional resistance. The coil resistance of the choke can also serve as a resistance. The combination of R and L establishes the filter characteristic. The formula for a Butterworth RLC half-section is as follows:

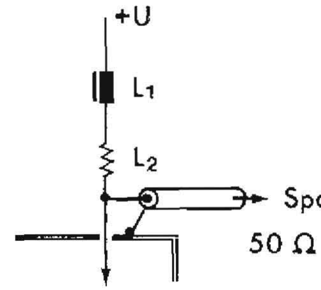


Fig.2: Measurement of interference levels from circuits
L1 = choke 10 ... 100µH
L2 = air-spaced choke 1µH

$$R = \frac{\sqrt{2}}{\omega C}; \; L = \frac{1}{\omega^2 C} \text{ with } \omega = 2 \cdot \pi \cdot f_{-3dB}$$

For ready-made chokes, the ohmic resistance is often specified. This should be somewhere near the calculated value, so as not to restrict the low pass in its function. The additional components are arranged inside and outside in the immediate vicinity of the feed-through capacitor, using short leads. Now anyone might maintain that filters in end stages manage without resistances. However, this is incorrect, since these filters do not become a low pass until a correct termination is provided (e.g. 50Ω impedance).

3. INTERFERENCE ON POWER SUPPLY LINES

In principle, high-frequency radiations are difficult to detect and usually require free field measurements using extensive measuring equipment. It is simpler to investi-

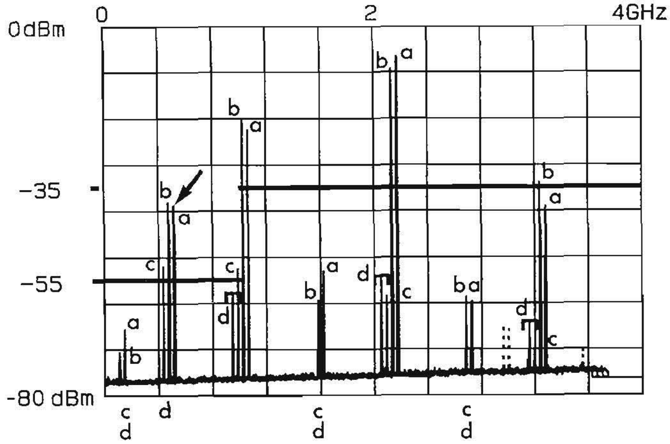


Fig.3: Measurement as per Fig.2 on an Oscillator Module

(a) Original condition of circuit
(b) Connection pin placed on earth side
(c) Position as for (b) but decoupling through 1nF feed-through capacitor
(d) Additional 1µH choke in housing
Arrow shows Television channel-30

gate individual circuits of a piece of equipment, for as long as a signal is linked to a circuit it can be measured in a defined way. For example, a measurement can show what power is generated by a signal linked at 50Ω.

Fig.2 shows a simple test rig. A power supply line is fed into a sheet-metal housing, the propagation of the high frequency to the exterior is prevented by filter choke, and the power emitted is fed through a coaxial cable to the spectrum analyser. An important factor here is a short and unambiguous earthing system for the coaxial cable, fitted directly to the lead-in wire. An additional coupling capacitor is superfluous, since most measuring equipment has AC coupling.

Measurement of the high-frequency current, using a suitable current clip, is also normal in the frequency range up to a few decades of MHz. This is especially true for circuits which are at a high potential, such as power supply lines. So that a defined current can flow, a clever terminal resistance is mounted at the end of the line to be investigated, consisting of a set of hinged ring cores, damping in the closed condition (absorption current clip).

The relevant standards lay down a limit of 30 MHz. Below this limit, we are assumed to be dealing with signals from circuits, with radiation appearing only above it.

4. PRACTICAL EXAMPLE OF SHIELDING

Unfortunately, EMC is an unknown term, even for many serious suppliers of assemblies and equipment. What other explanation could there be for the neglect in this area?

Here's an example.

I took a ready-to-operate 13cm oscillator module, ready to hand from my stock. The assembly was powered at operating voltage in accordance with Fig.2, and simultaneously measurements were carried out at the mains connection. The result can be seen in Fig.3 - a latticework fence of signals (a) of varying intensity! The level on the power supply line was so high that you could almost have powered a mixer from it. This condition is caused by the design, since the power supply connection is right next to the signal outputs. The fairly uniform spectrum marked as (b) was measured on the "cold side" of the board when the assembling bolt had been placed on the other side. The high level raises the possibility that the entire assembly is "hot". Spectrum (c) shows the behaviour after the incorporation of a feed-through capacitor - high frequencies are comparatively better dampened because the inductivity of the feed wires becomes noticeable. With an additional choke, only three very strongly dampened lines (d) remain visible.

The limiting values applying to interference from circuits vary with the type of equipment and licensing, particularly with regard to high-frequency inputs and outputs. Average values are determined so that the admissibility of the level can be estimated. Here they are -55dBm up to 1 GHz and -35dBm above that. The question remains what level of interference would continue to penetrate to the outside world once the assembly was integrated into a piece of equipment. In any case, in recognition of the assembly's enormous interference level, an FTC and a choke should be incorporated as simple but effective EMC measures. Among other things, there is spurious radiation at a high level of 544 MHz, i.e. precisely in television channel 30. This can naturally cause a lot of unnecessary irritation.

Thus the thesis is also disproved that overtones from radio amateurs' operations always "come home", that is onto amateur radio bands.

Of course, a screening box for such an assembly must be soldered up on all sides, so that the box can be prevented from acting like a slot antenna and emitting unwanted radiation.

5.
VARIOUS LOGIC FAMILIES

There are a large number of logic families in which every member is a strong source of interference. It's not so bad with the standard CMOS 4000 and 74C00 ranges. Modern logic families obtain a large part of the operating speed because they transfer smaller input capacitances, but the total

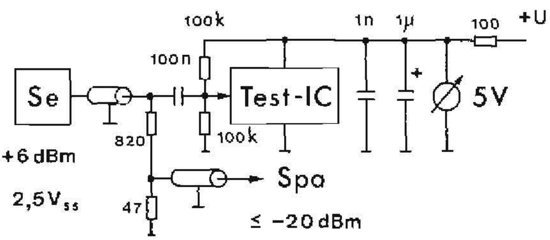


Fig.4: Interference measurement at Input of Digital Modules

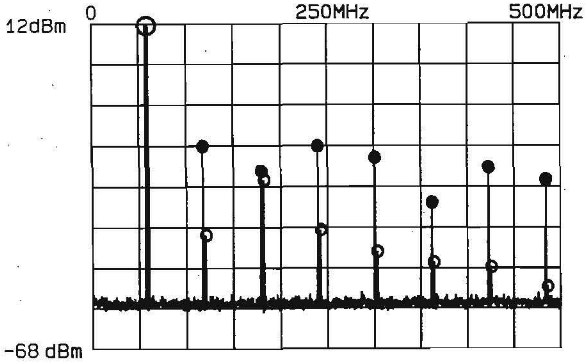


Fig.5: Measurement as per Fig.4 on two Gates at 60 MHz Dots: 74HC00 Circles: 74HC04U

interference occurring is greater. Printed circuit boards with digital circuits, like motherboards for PC's, belong in an HF-proof sheet-metal housing. Of course the computer companies who use processor clock frequencies in advertising and display them on the PC screen don't want to know about this.

Standard transistor-transistor-logic (TTL) modules, the modern 74S, 74LS, 74ALS, 74F ranges and the MOS equivalent types 74HCT and 74ACT should be avoided as far as possible. If speed of operation is required, all that's left is the 74HC range, plus (with reservations) the "harder" 74AC version. There are two important reasons for this. Firstly, HC modules operate at power supply voltages of 2 Volts and upwards, which makes for outstanding characteristics. The speed of operation increases only moderately. The high-frequency interference dispersion, by contrast, is strongly reduced. The second reason is that shielding measures frequently reduce the digital signal deviation. However, the subsequent gates cope with this without any alteration, whilst with TTL gates it can lead to problems, since their switching threshold is critical.

Experience teaches us that, with a power supply voltage of 5V, any models from the 74LS, 74HCT and 74HC ranges can be mixed together, so that it is possible to clean up existing circuits.

At the moment, components in processors are still operating at 5 Volts, but a 3.3V standard is already being drawn up. Thus the trend is actually towards lower volt-

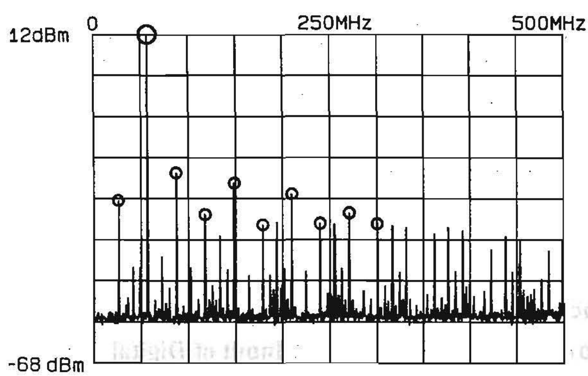


Fig.6: Measurement as per Fig.4 on a 74HC4040 Divider

ages, which is a consequence, among other things, of the enormous power consumption of lap-top computers to date.

The module outputs are very unclean, as might be expected, but what about the inputs? HC-MOS standard dividers are at the forefront, because of their high-resistance inputs. Fig.4 shows a suitable test rig. A signal of +6dBm is applied at the IC input through a standard signal generator, which creates a voltage dispersion of about 2.5Vss. The signal is sufficient to switch the IC cleanly and yet still low enough to generate no saturation and thus no additional overtones. A spectrum analyser is connected in parallel to the generator through a high-resistance voltage divider.

With this kind of measurement, clean de-coupling from the power supply voltage is important. This can best be obtained through an additional resistance in the power supply line. Unfortunately, the current consumption fluctuates very widely, depending on the operating frequency, so that the power supply has to be re-adjusted. The maximum power supply voltage must under no circumstances be exceeded when the generator is switched off.

Fig.5 shows the measurement result for two gates, one buffered (74HC00) and one unbuffered (74HC04U). The differences are plain to see. The input level of the IC can be calculated through the damping

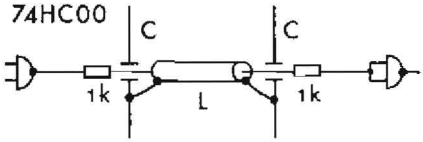


Fig.7:
Digital Transmission using Feed-through Capacitors and Coaxial Cables (100pF/m, e.g. RG58 or RG174; with C=100pF and L=1m
Delay 0.2 to 0.3µs, DC up to 5 MHz

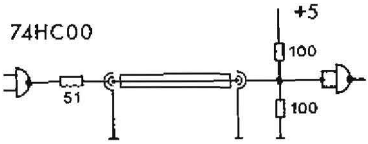


Fig.7:
Digital Transmission using Coaxial Plugs and Sockets and Cables
10m RG58: DC up to 70 MHz
50m RG58: DC up to 25 MHz

factor of the divider. The diagram has been scaled accordingly. Opinions may be divided as to whether the value of "12dBm" is correct for the generator at no-load. But for the interference level the scale is certainly correct, since precisely this value is generated at the internal resistance of the generator. In any case, it is clear that even the interference signals from the gate inputs, judged by the criteria of communications technology, are at very high levels!

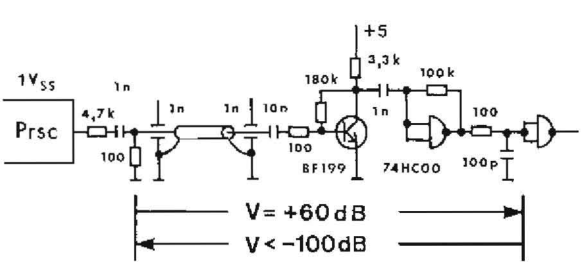


Fig.9:
Transmission between sub-assemblies using the example of a Pre-divider
Specifications at 1MHz
Operating range 30 kHz to 10 MHz

Fig.6 shows the measurement result for a 74HC4040 divider. Here the sub-harmonics are especially striking, especially the frequency divided by 2. Both the module's internal functioning and its pin grouping may contribute to this. The data book gives input capacitances of a few pF, so they are not earthed at all but are, in the worst case, wired directly to the output of the module.

6. DIGITAL SIGNAL TRANSMISSION

Digital system interfaces are a significant source of interference. With suitable screening measures, the interference level can be considerably reduced. In addition to the use of "digital plugs" with built-in low passes, something like D-plugs, something can be achieved using ferrite components and little tricks. Thus flat cables can be made 1m longer and coiled into air-core coils. Even better are shielded round cables with well-defined earthing systems on both plugs.

Serial interfaces are preferable, since here the screening measures can be concentrated on a few conductors and the transmission rate can be set to reduce interference if necessary. Moreover, the step to fibre optic transmission is simpler with serial interfaces. This may be necessary both for interference suppression and for de-coupling (e.g. in the medical field).

It is not always necessary to use expensive plugs to ensure clean, low-radiation transfer. Feed-through capacitors will suffice as well.

Fig.7 shows a possible circuit for low-frequency digital signals. In this case, the cable can be comprehended as a capacity which is switched in parallel to the feed-through capacitors. The resistances and the screen earthing should be connected in close vicinity to the lead-in wire. This simple circuit produces running times which are worth taking into account. The circuit shown in Fig.8 is suitable for faster digital signals. The cable is terminated in its own impedance on both sides without any displacement of the switching threshold on the receiver side. With longer cables, the cable damping becomes noticeable, which means that the subsequent gate no longer has precise advance control. Should the driving gate be over-loaded at 100, several gates from the same housing can be wired up in parallel.

Fig.9 shows a further example. Here a UHF pre-divider is linked to a digital section (PLL module). The oscillator and the pre-divider are in a joint housing, with the digital section in a separate housing of its own.

(To be continued)

MATERIAL PRICE LIST OF EQUIPMENT

described in VHF COMMUNICATIONS

DB1NV	**Digital Image-Store for the Spectrum Analyser**	**Art. No.**	**Ed. 4/1991**
PCB	DB1NV010	6477	DM 44.00
Components	Processor P80C31; 12 ICs; 1 Reg. IC, Transistor; Zener Diodes; Silicon Diodes; Chokes; RAM; EPROM DB1NV 010; 4 x 2k Trimpots; Crystal	6478	DM 276.00

DB1NV	**Tracking Generator for the Spectrum Analyser**	**Art. No.**	**Ed. 1/1992**
PCB	DB1NV011	6479	DM 31.00

F6ILR F6BXC	**A Digital Slow-Scan Television Transmit Coder**	**Art. No.**	**ED. 3/1992**
PCB	SSTVCODE1 (KM Publications)	SSTV1	£ 28.50

DB1NV	**Broadband VCO's using Microstrip Techiniques**	**Art. No.**	**Ed. 4/1992**
PCB	DB1NV012	6480	DM 33.00
PCB	DB1NV013	6481	DM 33.00
Components	400 - 1250 MHz 3 x BB619; 1 x BB811; 1 x BFG96; 2 x AT42085; 1 x BFQ69; 2 x 2.2nF & 1 x 27pF Feed-through Cap.; SMC Connectors; 2 x 0.47 H SMD Choke; 1 x housing 74x55x30mm; 1 PCB DB1NV 012	6482	DM 81.00
Components	450 - 1450 MHz as above but: 1 x BFG65 instead of BFG96	6483	DM 81.00
Components	800 - 1900 MHz as above but: 4 x BB811 instead of BB619 and PCB DB1NV 013 instead of 012	6484	DM 85.00

Tuning Diode	BB619	10 off	10450	DM 20.00
Tuning Diode	BB811	10 off	10451	DM 31.50

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