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DIGITAL PROCESSING OF CTFM SONAR SIGNALS.

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1. INTRODUCTION

Traditionally most sonars have been based on pulse-echo technology. This is a consequence of the relative ease with which this type of signal can be generated and analysed. Although the relative merits of FM (frequency modulated) sonars have been recognised for over 50 years, the available technology has made the implementation of FM sonars difficult. These difficulties are now being overcome with the advent of digital frequency synthesisers and digital spectral analysis techniques.

There has been a long history of CTFM sonar development at the University of Canterbury, including such diverse applications as blind aids, heart monitors (cardiophone), fish finding sonars, diver's sonars, robotics and underwater classification sonars [1—3]. The block diagram of a simple analogue CTFM sonar is shown in Fig. 1. This paper looks at the properties of the CTFM signal and how it can be generated, demodulated and analysed by digital means.

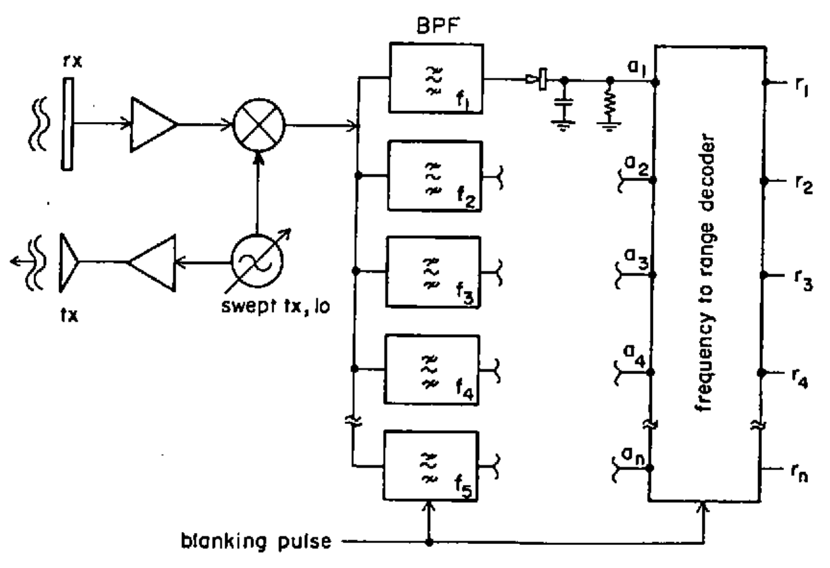

Fig 1: Simple analogue CTFM sonar processing

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2. THE CTFM SIGNAL

A CTFM signal can be considered to be a coherent train of FM pulses where the duration T_p of each pulse is equal to the pulse repetition period T. This period T is generally known as the sweep period. Note unlike chirped sonars, the duty cycle of CTFM sonars approaches unity.
The most common CTFM signal is based on a linear FM pulse of the form

$$s_p(t) = \mathrm{rect}\left(\frac{t - T/2}{T}\right) e^{j2\pi(f_0 t + \frac{1}{2}\mu t^2)} \tag{1}$$

where f_0 is the initial frequency, μ is the sweep rate, and T is the sweep period. The sweep rate is proportional to the slope of the instantaneous signal frequency and in terms of the terminal and initial sweep frequencies is

$$\mu = \frac{f_1 - f_0}{T} \tag{2}$$

Note that for a decreasing frequency sweep (down-sweep), the sweep rate is defined to be negative. For example, $\mu = -f_0/2T$ for a one octave, down-sweep. The linear FM pulse defined by Eqn.(1) is repeated every T seconds to form a continuous signal.

2.1 CTFM signal generation.

The CTFM signal is traditionally generated by applying a sawtooth voltage ramp to a Voltage Controlled Oscillator (VCO). Typically the voltage ramp is generated by integrating the output of a constant current source; the greater the current, the faster the sweep rate μ. At the start of each sweep, the integrator is reset to a value corresponding to the initial frequency. It was noted at an early stage in the development of CTFM sonars that the performance of the sonar was limited by the linearity and stability of the frequency modulated oscillator. Extremely high linearity of the frequency sweep is required, especially for long sweep periods, because any non-linearity causes an error in the demodulated difference frequency.

2.1.1 Digital Generation of the Transmitted Waveform. The introduction of digital signal generation methods to CTFM sonars, allowed high linearity and stability of the frequency modulated oscillator for the first time. Zehner [4] patented the method of producing a very linear voltage ramp, using a counter followed by a digital to analogue converter (DAC) to generate a staircase voltage waveform. This ramp voltage was used to linearly sweep a VCO as before. However, the VCO output was used to clock a counter which successively addressed the samples of a sinewave stored in memory (usually an EPROM). From these samples, a swept linear FM signal was reconstructed using another DAC. The number of samples required depends on the time-bandwidth product of the signal and the oversampling factor used. The sampling theorem suggests that at least two samples are needed per cycle but to ease the specification of the post DAC filter oversampling is used. This filter removes the unwanted harmonics in the spectrum of the sampled signal. For a sweep of period T, maximum frequency f_m, and an oversampling factor γ, then $2\gamma f_m T$ samples required to store the entire sweep. This number can be reduced by storing only the baseband of the sweep in which case only $2\gamma BT$ samples are needed, where B is the signal bandwidth. Multiplying the baseband signal by a carrier signal, and removing the unwanted sideband by filtering, generates the desired sweep for transmission. However the filtering operation is difficult and it is often better to generate a single sideband using quadrature techniques . Twice as many samples are now required to store both the in-phase and the quadrature phase reference signal, but often this is advantageous since both are often required for the demodulation process as explained in later sections.

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3. THE CTFM ECHO SIGNAL

To simplify the following explanation, it is assumed that there is a single stationary point target in a homogeneous medium. (The effect of a moving target on a CTFM signal is omitted, as is any frequency dependent amplitude and phase fluctuations.) The echo signal is thus just a replica of the transmitted signal, delayed by $\tau = 2r_a/c$,

$$e(t,\tau) = \mathbf{A}s(t-\tau) \tag{3}$$

where the complex scale factor $\mathbf{A} = Ae^{j\phi}$ describes the target reflection properties and the propagation losses due to spreading and absorption. Note that the notation $e(t,\tau)$ describes a single echo delayed by τ. The general echo signal comprises the sum of all the echoes. The echo signal has a frequency discontinuity at $t = nT + \tau$, whereas the transmitted signal has a frequency discontinuity at $t = nT$. These frequency discontinuities often generate transients that can mask weaker signals. Usually the largest transient is caused by the transmitted signal flyback, and hence it is desirable to momentarily blank the output of the demodulator just after the start of each transmitted sweep. The flybacks in the echo signal are range dependent, however, and therefore cannot be blanked as easily. Moreover the transients generated by the echo signal flybacks are usually too small to be of consequence, excluding of course echoes form large close-range targets.

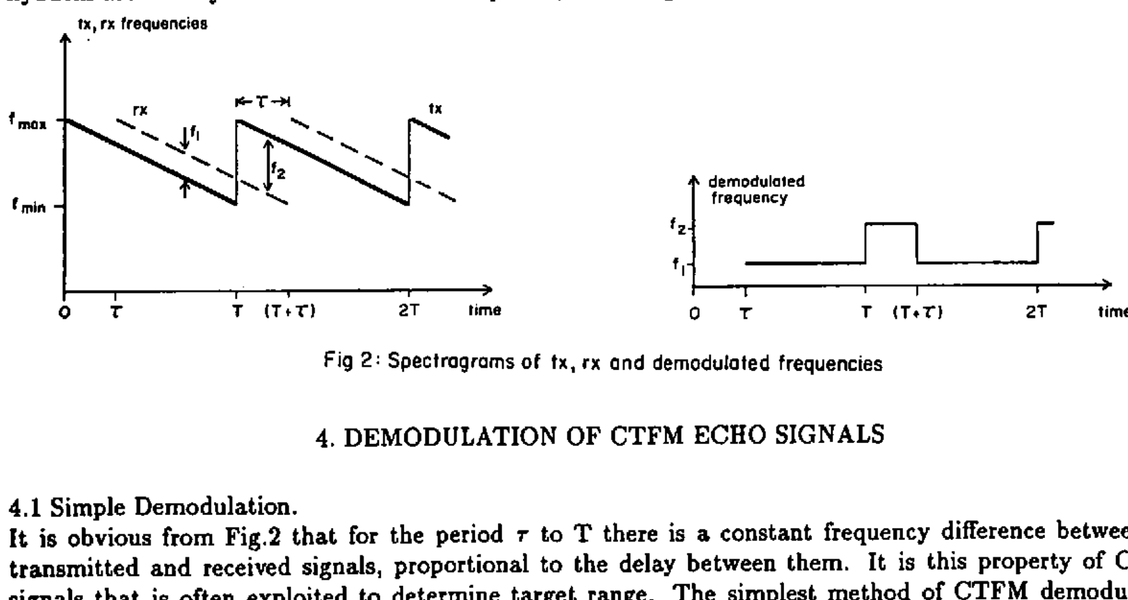

Fig 2: Spectrograms of tx, rx and demodulated frequencies

4. DEMODULATION OF CTFM ECHO SIGNALS

4.1 Simple Demodulation.

It is obvious from Fig.2 that for the period τ to T there is a constant frequency difference between the transmitted and received signals, proportional to the delay between them. It is this property of CTFM signals that is often exploited to determine target range. The simplest method of CTFM demodulation —sometimes referred to as *heterodyne correlation* or as the *spectrum analyser technique* — relies on the fact that there is a substantial overlap between the transmitted and received signals. The received signal is multiplied directly with the complex conjugate of the transmitted signal to obtain the desired difference frequency signal. However, it is difficult in practice to generate the complex echo signal from the real echo, since this requires the use of wideband Hilbert transformers. So instead, the real echo signal is multiplied with the real transmitted signal. This produces the desired difference-frequency component and in addition a sum-frequency component around $2f_0$ which is removed by a low pass filter (LPF).

This simple demodulation operation may be expressed concisely as

$$d(t) = s(t)e(t) \odot i(t) \tag{4}$$

where $i(t)$ is the impulse response of the low pass filter and $\odot$ is a convolution process. The instantaneous demodulated frequency is written as

$$f_a(t,\tau) = \begin{cases} |-\mu\tau| & = 2|\mu|R/c, \quad \tau \le t \le T \\ |\mu(T-\tau)| & = |\mu(2R/c - T)|, \quad T \le t \le T+\tau \end{cases} \tag{5}$$

and in the interval $\tau \le t \le T$, is directly proportional to the target range as expected. So for a measured difference-frequency f_a, the indicated range r_a is given by

$$R_a = \frac{cf_a}{2|\mu|} \tag{6}$$

However for the interval $0 \le t_s \le \tau$ after the start of each transmitted sweep, the range is not directly related to the difference-frequency. This occurs when the received echo from the previous sweep is demodulated by the current transmitted sweep, and is described by the second term of Eqn.5. Since the simple demodulation technique cannot differentiate between positive and negative frequency differences, a frequency-to-range ambiguity exists and the interval when the difference-frequency is not proportional to the target range is usually ignored by blanking the demodulator output for a period T_b (called the ***blind time*** or ***lost time***) after the start of each transmitted sweep. The longer the blanking interval, the greater the unambiguous range $cT_b/2$. However a long blanking interval is undesirable, because both the signal to noise ratio and range resolution are degraded by a factor $\beta = T_b/T$. Therefore, as a compromise between blind time and maximum unambiguous range, the blind time is typically limited to 10–30% of the sweep period.
There are a number of other disadvantages to the simple method of demodulation described by Eqns 4,5. Obviously as the blanking interval is increased, the quality of the constant tone τ to T degrades. This is not so much of a problem for long sweep periods but for short sweep periods, this unwanted modulation may mask the target sounds. Similarly this discontinuous tone may cause unwanted transients when applied to a spectrum analyser comprised of banks of highly selective band-pass filters.

4.2 Quadrature Demodulation.
Although it is difficult to produce $\hat{e}(t)$ from $e(t)$, it is relatively easy to generate both $s(t)$ and $\hat{s}(t)$ using digital synthesis techniques (the 'hat' denotes a Hilbert transform). By multiplying the received echoes with the in-phase and quadrature phase components of the transmitted signal, it is now possible to differentiate between positive and negative frequency differences — i.e. whether the echoes are higher or lower in frequency than the transmitted signal. Fig. 3 shows this in some block diagram form where the analogue bank of filters has been replaced by a phase sensitive digital FFT as well as an analogue quadrature demodulator. In this system, echoes from the previously transmitted sweep can be distinguished from the echoes of currently transmitted sweep. With a down-sweep, for example, the echoes from the previously transmitted sweep are always lower in frequency than that of the currently transmitted signal, and echoes from the currently transmitted sweep are always higher in frequency than that of the currently transmitted signal as shown in Fig. 4(a). Since the positive and the negative frequencies now have different frequency-to-range decoding, some form of phase sensitive spectrum analyser is required to ensure that the correct form of decoder is used for the positive and negative frequencies. Fig.4(b) shows the demodulated frequencies for a single target at a constant range.

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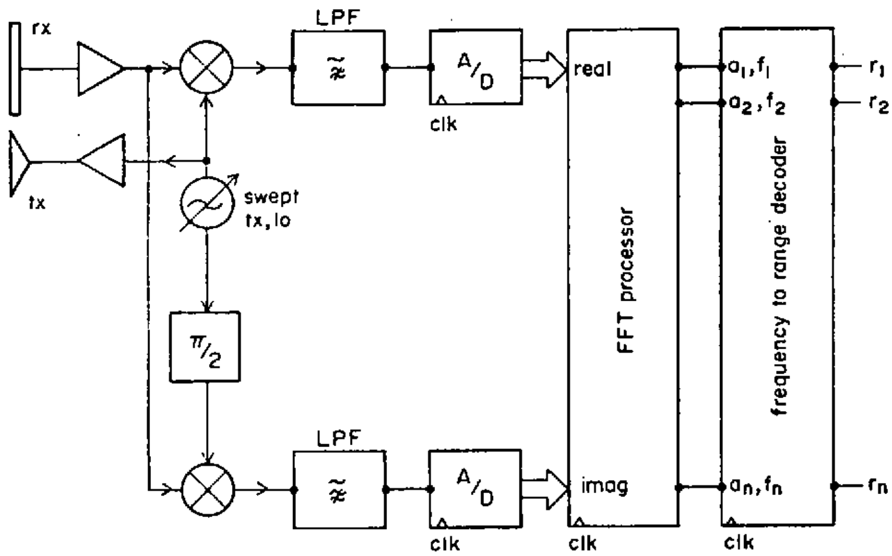
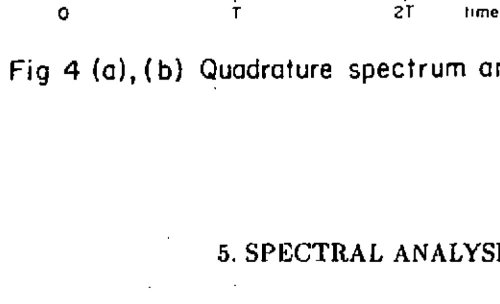


Fig 3: Quadrature analogue demodulation with digital spectrum and frequency to range decoding



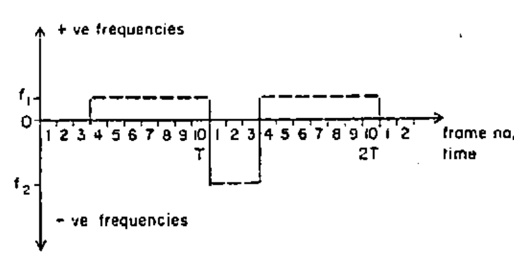


Fig 4 (a), (b) Quadrature spectrum analysis

5. SPECTRAL ANALYSIS OF DEMODULATED CTFM SIGNALS

For single target range measurement (e.g. altimeters), a simple phase locked loop or zero-crossing counting technique can be used to determine the target range. If there are multiple targets present in the sonar beam, the demodulated signal contains a number of difference frequencies. Provided the system is linear, there is a frequency component corresponding to each target, and thus the combined demodulated signal is

$$d(t) = \sum_N A_n \cos(2\pi f_n t + \phi_n) \tag{7}$$

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To separate the signal components and to sort them into discrete range bins, it is necessary to use spectral analysis techniques. (This is in contrast to pulsed sonars that separate the target echoes in the time-domain.) The spectrum from a multi-target CTFM sonar is usually obtained by using banks of bandpass filters, where the number of filters needed depends on the desired range comprehension and range bin width. Obviously the better the filter selectivity, the greater the number of filters that are required to cover a given range interval (bin). If the phase of each range bin is required, it is necessary to use a quadrature demodulation technique with an additional filter bank for the quadrature signals. The output from the two banks of filters (i.e. the in-phase and quadrature-phase banks) are then fed to a bank of phase detectors, or are time multiplexed between a single phase detector. A bank of N filters provides N independently varying outputs each of which is an analogue signal being the sum of all echo structures arising from reflection and scattering within the annulus corresponding to the frequency band of the filter.
The compressed CTFM sweep has the shape of a sinc function, with a 3dB width of $0.88/T$, corresponding to a nominal range resolution of

$$\Delta R = 0.88 \frac{c}{2|\mu|T} \tag{8}$$

Note that the range resolution of a CTFM depends on the frequency resolution Δf of the spectrum analyser, with the highest resolution obtained when $\Delta f \sim 1/T$. However this range resolution is not always needed, in which case the analysis bandwidth can be covered with fewer bins. Moreover since the filters are now less selective, they respond more rapidly. Thus there is a trade-off between range resolution for speed of response; something no pulsed sonar can match [2,3]. In addition, it is now possible to make a number of range measurements within the sweep period, whereas a pulsed sonar is limited to one look per pulse.

5.1 Digital Spectral Analysis.
Most digital spectral analysis is based upon the Discrete Fourier Transform (DFT). In simplistic terms, the DFT can be considered similar to a bank of parallel, constant bandwidth, bandpass filters. The two most common implementations of the DFT being the Chirp Z-Transform (CZT) and the Fast Fourier Transform (FFT). The FFT is just a numerically efficient algorithm for calculating the DFT and both special, and general, purpose array processors designed for block FFT's are now quite common for input sample lengths of 128, 256, 512 or 1024 time samples. A block FFT may be computed over the entire echo lasting T sec or alternatively, a series of time-contiguous smaller block FFT's may be computed. One disadvantage of taking a single block FFT over the entire sweep period (other than the limitation of the array processor) is that the flybacks in the echo signals occur at times in the sweep depending on the target ranges. Taking a FFT over a flyback produces a 'glitch', or in the worst case, complete cancellation. To overcome this problem, a series of shorter length, but contiguous, FFTs may be computed each one covering a different part of the sweep. These 'sub-FFTs' are then combined into an effectively longer FFT, but one that doesn't encompass the 'glitch'.

6. DIGITAL DEMODULATION AND ANALYSIS

Once the bank-of-filters spectrum analyser (Fig. 1) has been replaced by a phase sensitive block FFT process (Fig. 3), much of the demodulation can be converted to a purely digital process. After the preamplifier (the only non-digital stage), the signal is split into in-phase and quadrature-phase channels and sampled at an appropriate rate. At first glance, it would seem that a sampling frequency f_s of just more than twice the highest transmitted frequency would suffice. Later on we show that this is not enough, and that sampling rates need to be considerably higher. After the A/D (which in fact may be shared by the two channels), the two signals are multiplied by a cosine waveform for the in-phase and by a sine waveform for the quadrature phase channels. At the outputs of the multipliers the combined sum- and difference-frequencies are LPFed

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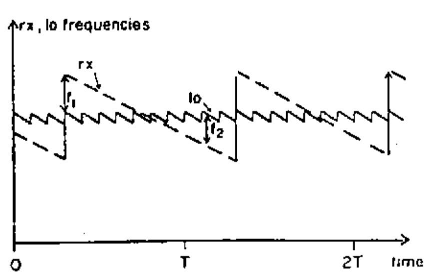
to leave only the difference-frequency signals in each channel. However for the output of the multiplier to be free of aliased terms, the sampling rate used by the multiplier must be at least twice the *maximum sum-frequencies*. This means that the sampling rate of the input signals must be very much higher than twice the maximum transmitted frequency as originally suggested. After the LPF has removed all the sum-frequencies, the remaining signal can be decimated down to a sampling frequency just more than twice the maximum difference-frequency; a considerable saving in computational requirements. The post decimation samples that now represent only the difference-frequencies are sent to the real and imaginary inputs of the spectrum analyser. The outputs of the analyser are then used as input to the appropriate frequency-to-range decoder. In practice it is usual for a single array processor to do the multiplication, digital LPF, the spectrum analysis, and finally the frequency-to-range decoding. Each block of input samples are cos- and sine- premultiplied, filtered and decimated before the FFT algorithm is applied. This is a relatively inefficient process since each block FFT needs a different set of cos- and sine- premultiplication coefficients which must be loaded into the fast memory of the array processor (usually by DMA). A faster, and consequently, more efficient process might be to use the same cos- and sine- premultiplication coefficients permanently located in some part of array processor fast memory. Every block or frame of input samples would then be demodulated by an identical LO covering the same restricted frequency band as shown in Fig. 5(a). The resultant difference-frequencies (Fig. 5(b)) now shows a stepped, piecewise continuous, dependence for a target at a constant range. The disadvantage of this system is that the frequency-to-range decoding becomes more complicated but the saving in DMA and so computation time might be worth it.



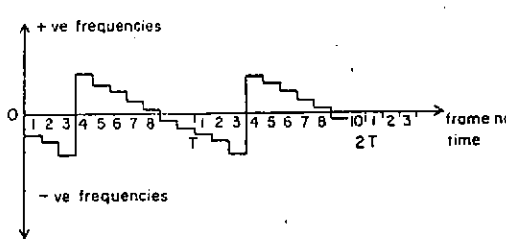


Fig 5 (a), (b) Restricted local oscillator with quadrature spectrum analysis

7. QUASI-DIGITAL DEMODULATION AND ANALYSIS

The limitation to a purely digital CTFM sonar is the necessity to sample the input signal at twice the frequency of the highest sum-frequencies expected at the output of the digital multiplier. A possible solution to this limitation is to convert all the echo frequencies down to the lowest possible baseband frequency while still preserving their phase relationship. This can be done with a fixed frequency down-convertor just before the A/D convertor as shown in Fig. 6. (The down convertor comprises a fixed frequency LO as one input to an analogue multiplier, followed by an analogue LPF at the output). The frequency spectrographs of the echo signals after downconversion and of the subsequent cos-and sine- premultipliers are shown in Fig. 7(a). As with the purely digital CTFM demodulation technique, a restricted bandwidth cos- and sine- premultiplication can result in a more efficient computational process since the multiplication coefficients are loaded into the fast memory of the array processor only once. Frequency spectrographs of the down-converted echoes and of the cos-and sine-premultipliers are shown in Fig. 7(b).

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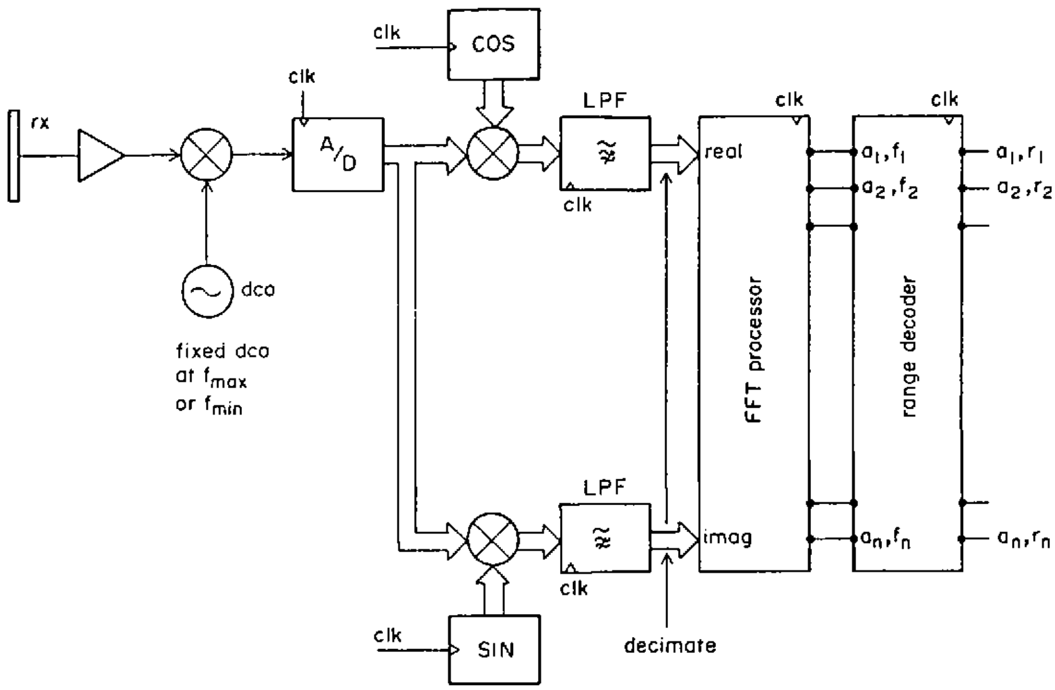

Fig 6: Hybrid analogue/digital CTFM processor with fixed down-converter (dco)
Master clock controls all devices: tx oscillator, fixed lo, A/D, cos and sin LUTs, digital LPF, FFT, range decoder

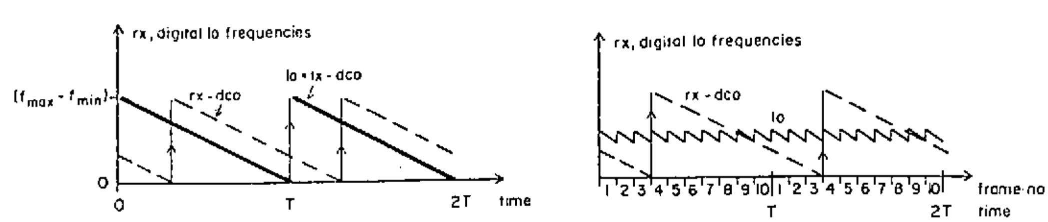

Fig 7: Baseband implementation of CTFM processing
after fixed downconverter
(a) with full baseband lo (b) with restricted lo

8. REFERENCES

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