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The characteristics required of a microwave frequency agile source suitable for modern, coherent, multi-mode radar systems are discussed. Several solutions to the problem of generating a sufficiently large number of channels with minimum compromise to spectral purity are examined. Some factors relating to the hardware design of a suitable frequency synthesiser are considered, including a brief review of the current performance capability of two of the more important system components. The paper concludes by comparing some published results, and indicating some likely future trends.

Introduction

Frequency agility is an essential feature of all modern radar systems intended for military applications. In addition to the obvious advantage of increasing resistance to electronic counter-measures, agility is of benefit in avoiding prolonged fades due to target scintillation, in suppressing unwanted echoes from previous pulses, and in providing a useful measure of sea clutter decorrelation. The ability to operate on a number of different transmission and reception frequencies also offers the possibility of interrogating transponders, communicating with missiles or other installations, and, in the case of coherent systems, achieving high resolution or imaging modes.

The first frequency agile radars appeared in the mid-1960s, following the development of the spin-tuned magnetron. Operating on a pulse to pulse agility basis, these systems using spin-tuned and later types of agile magnetron continue to supply an efficient solution to the provision of frequency agility where only non-coherent radar modes are required.

Whereas it is advantageous for non-coherent radars to change frequency from pulse to pulse, coherent systems by their nature are generally restricted to pulse burst to burst agility. Although modern developments in magnetron technology such as injection-locking have made it possible for frequency agile tubes to operate coherently, most high performance systems utilise an amplifying device as the transmitter. To achieve the required power level and agility bandwidth, this device is normally a travelling wave tube, in the case of current generation systems based on a passive antenna.

Fig. 1 shows an example of a typical architecture for the microwave and radio-frequency section of a system of this type. Since travelling wave tubes have much lower peak power ratings than magnetrons of equivalent mean power, it is usually necessary to transmit relatively long duration pulses, and recover range resolution by pulse compression techniques. This demands frequency or phase modulation of the transmitted waveform, which is normally carried out at a relatively low intermediate frequency, such as 60 MHz. Because the agility bandwidth will be much higher than this figure, two stages of up-conversion are necessary to translate the modulated waveform up to the transmission frequency. Following transmission, the received signal is down-converted to the frequency at which the original waveform was generated, and then further translated into in-phase and quadrature components at baseband before digitisation and subsequent signal processing.

Compared to even a non-coherent magnetron system, implementation of the microwave frequency agile local oscillator sub-system appears at first sight to be deceptively simple, since there is no longer any requirement for a

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complex automatic frequency control system to lock the local oscillator to an IF offset away from the magnetron frequency immediately after each transmission. The situation seems further simplified by the fact that the system architecture described allows frequency modulation of the transmitted waveform, necessary for pulse compression or fm ranging, to be carried out independently from generation of the microwave frequency agile signal. Similarly, clutter tracking, often necessary in airborne systems, can be implemented via one of the fixed frequency LOs, rather than by detuning the microwave source. Despite these apparent simplifying factors, however, the design of the microwave frequency agile source for a coherent system remains an exceedingly difficult problem. This arises because of the spectral purity requirement imposed by modern radar specifications.

### Spectral Purity Requirement

The spectral purity required of the microwave frequency agile source must be considered in terms of phase noise, close to carrier discrete spurious signals, far from carrier discrete spurious signals, and spurious modulation and break-through arising from environmental effects.

### Phase Noise

The phase noise contribution of the various building blocks in the system must be maintained sufficiently low that the overall radar meets its sub-clutter visibility specification. This will depend on the prevailing clutter to noise ratio, and on the minimum signal to be detected. Since the frequency agile source normally operates at a microwave frequency at least ten times higher than any of the other local oscillators shown in Fig. 1, it is normally the most predominant source of oscillator induced phase noise. In general, its phase noise will also substantially exceed the added noise arising from the various amplifiers in the system, including the transmitter.

In the case of a pulsed doppler radar it is not strictly correct to specify the required source phase noise curve in the form of a hard and fast boundary. The reason for this is shown in Fig. 2. Since the noise spectrum associated with each spectral line overlaps into the identical spectra of other lines, the total noise power appearing in a given doppler filter will be the sum of the noise contributions from all spectral lines, weighted by the pulse spectral envelope and IF filter function.

### Close to Carrier Discrete Spurious Signals

In the case of modulation sidebands due to phase noise, the total unwanted power at the output of a given doppler filter is found by integrating the noise spectral density over the filter bandwidth. This does not however apply to discrete spurious signals, where the peak level will appear at the output of the filter. In the absence of a significant noise contribution from adjacent lines in a pulse spectrum then, the maximum level of close to carrier discrete spectral components is given by offsetting the relevant maximum phase noise curve by a positive amount equal to the doppler filter bandwidth relative to 1 Hz. For a phase noise limit of  $-125$  dBc/Hz at 10 kHz offset and a typical doppler filter bandwidth of 1 kHz for example, the maximum level of a spur at 10 kHz from carrier would be  $-95$  dBc.

### Far From Carrier Discrete Spurious Signals

As we have seen, it is a simple matter to specify the maximum permitted levels of close to carrier spurious signals once the phase noise boundary and doppler bandwidths are known. The situation with far from carrier spurious signals is somewhat more complicated. In order to control both phase noise and long-term stability, the microwave frequency agile source is invariably realised by some form of frequency synthesiser. Such systems generally result in far from carrier spurious signals tending to occur at offsets from carrier equal to harmonics of the channel spacing, and sometimes also at sub-harmonics of this spacing.

Spurious signals within the doppler band at the receiver output may then be generated by mixing products being formed between the frequency agile local oscillator channel selected, the spurious lines surrounding this channel, the second LO and its harmonics, the third LO and its harmonics, and the doppler shifted signal passing through the receiver. An obvious first step towards the solution to this problem would seem to be to derive all frequencies employed in the entire system, including the prf, from a single master reference oscillator. Such a move however is not always practical, since it would impose severe restrictions in the choice of channel frequencies, unless changes to the IFs were acceptable. In addition, in the case of an airborne system, it is often desirable to shift the mainbeam clutter spectrum to zero doppler by tuning the receiver centre frequency. This would be excluded by a requirement to derive all LO frequencies from a single master reference source.

In systems where the requirement is for only a small number of frequency channels, it may be possible to make significant economies in the design of the frequency agile source by choosing all frequencies such that the effect of any far from carrier spurious signals is to generate only mixing products lying outside the doppler band. This approach is very inflexible, however, and in general a level of around -80 or -90 dBc is imposed over at least the agility bandwidth, to ensure freedom from trouble under all circumstances. Achieving such a level together with the required phase noise performance is one of the most difficult problems in the design of the exciter for a coherent system.

#### Environmental Effects

Vibration, acoustic noise, and radio-frequency interference impose a severe threat to phase noise and close to carrier spurious signal performance. Although the effects of RFI may be reduced to negligible proportions by good engineering design, high levels of acoustic noise, and particularly vibration, are much more difficult to deal with. The basic reference signal in any synthesiser system invariably relies on either the piezo-electric effect, or on the dimensions of some microwave resonator structure, both of which are sensitive to mechanical stress.

All anti-vibration mounting systems are resonant structures, amplifying vibration levels at resonance, and therefore must be designed such that the vibration spectrum to be encountered falls at all times on the high side of resonance. This often results in a structure of considerable bulk. Anti-vibration mountings have no effect in suppressing acoustic noise, resulting in the need for enclosures lined with suitable absorbent material. It is worth noting that active vibration cancellation systems exist, and may find practical application in cases where the applied vibration is essentially confined to one axis. System specifications may impose both sinusoidal and random vibration requirements.

In designing or choosing the system reference source, it must be constantly borne in mind that the resonator technology offering the best phase noise performance under laboratory conditions may not also be the optimum choice for low sensitivity to vibration.

#### Number of Channels

The level of difficulty incurred in achieving the specified spectral purity performance in the absence of environmental factors is very much a function of the number of frequency channels required. In general, a step function in complexity arises when the number of channels and the agility bandwidth are such that the channel spacing becomes within an order of magnitude of the offset frequency range over which the phase noise must follow that of some reference signal. Considering, for example, an agility bandwidth of typically 500 MHz at X-band, this occurs at upwards of around 10 channels.

Recent literature [1,2] suggests that current radar system requirements are for between 10 and 128 channels, according to the exact nature of the system. In military systems, the need to conceal knowledge of all channel frequencies available is of obvious importance. This may be achieved by avoiding transmission of some channels in normal operation. However, if all channels are derived from a single reference oscillator, then knowledge of only a few channel frequencies enables the probable locations of the others to be predicted with great accuracy. This is a severe disadvantage of most frequency synthesis techniques, which inherently provide regularly spaced channels. Derivation of channels from more than one reference oscillator is therefore attractive, not only from a tactical point of view, but also in simplifying the synthesis scheme. Interrogation of transponders, or use of the system in a communications role, may also dictate the need for dedicated reference sources at specific frequencies.

### Channel Switching Speed

Much literature on the subject of coherent radar synthesisers suggests a need for channel switching in 10 microseconds or less to full phase stability. The need for such performance in doppler systems transmitting pulse bursts with many pulses on the same frequency is unclear, however, unless some form of interleaving of coherent modes on different frequencies is envisaged. Even in a 100 kHz high prf mode, for example, the interpulse time itself is as long as 10 microseconds, and a typical burst might contain several hundred pulses. One legitimate reason for demanding fast switching speed can be that many coherent systems are in fact multi-mode, and are required to operate non-coherently with pulse to pulse agility, as well as coherently with burst to burst agility. If pulse compression is employed in such modes [3] then close control of the microwave source frequency must be maintained in order to prevent firstly distortion of the encoded waveform, and secondly movement of the receiver centre frequency during reception, both of which would degrade the compressed pulse time sidelobe level.

### Some Frequency Synthesis Schemes

Compared to frequency synthesisers for other purposes such as communications and laboratory test gear, coherent radar synthesisers are characterised by the need for relatively few, widely spaced, channels, very high spectral purity, and relatively fast switching speeds. Because of this, many techniques commonplace in communications and other synthesisers are unsuitable for radar applications. In the following paragraphs several approaches to the radar synthesis problem are reviewed, assuming that the requirement is for between 10 and 128 channels within a 500 MHz agility bandwidth at X-band.

#### (a) Combined Reference Signal Multiplication and Channel Generation

All schemes in this category involve the formation of a channel set from harmonics of the reference signal falling within the agility bandwidth. This may be achieved either by means of a sampling phase locked loop, or by selecting the desired harmonics from a pre-filtered section of a comb spectrum by some form of tuneable filter.

The sampling loop approach has been described by Galini et al [4] and others, and is shown in its basic form in Fig. 3. This technique is attractive in its efficient use of hardware, and, if properly engineered, should give little degradation to the phase noise of the ideally multiplied reference. However, for systems requiring the ultimate in phase noise performance, the basic sampling loop becomes severely limited in its channel capacity. Best phase noise at X-band normally demands either a bulk wave crystal reference oscillator in the 100 to 150 MHz range, or a surface acoustic wave resonator oscillator at an even higher frequency, restricting the number of channels to around 5 or 6 within a 500 MHz bandwidth.

The number of channels may of course be increased by the use of two or more reference oscillators to provide interleaved comb lines. In this case it should be observed that the channel spacing is non-uniform, progressively decreasing towards the higher end of the agility band.

Other methods of increasing the number of channels include division of the reference frequency, and sampling a harmonic of the VCO output. Added noise data for ECL dividers shows that digital division of the reference by this technology will introduce unacceptable degradation to phase noise performance, although the divider noise cancellation technique described by Huffman [5] could alleviate this problem. However, even if such an approach is successful, division of the reference is restricted by the need to maintain a sufficiently high sampling rate for the desired loop bandwidth to be achieved.

Sampling a harmonic of the VCO output offers another approach to increasing the number of channels available. By sampling the second harmonic, for example, the number of channels is doubled. This technique has attractions not only in increasing the number of channels, but also in noise reduction possibilities. Implementation of the sampling phase detector however becomes more difficult as the sampling frequency increases.

Before leaving the sampling loop approach, it is worth noting that Amati et al [6] warn that considerable difficulty may be experienced in preventing the very broad spectrum of the high amplitude sampling pulse from giving rise to far from carrier spurious signals on the VCO output.

As an alternative to the sampling loop, some form of electronically tuneable filter may be employed to select the desired harmonic of the reference frequency comb. In this case, only that section of the comb spectrum following within the agility bandwidth need be presented to the tuneable filter. Tuneable filters are discussed further under category (e) below.

#### (b) Offset Loop Approach

Fig. 4 shows the principle of the offset loop approach in its most fundamental form. Starting with a reference signal at some arbitrary frequency, this is multiplied up to the agility band centre, and symmetrical pairs of channels formed by phase locking a microwave VCO to directly synthesised offsets from this signal.

Owing to their relatively low frequency compared to the microwave output signal, generation of the offsets may involve digital division of the reference signal without incurring a phase noise penalty.

Such a simple approach however is only satisfactory if the channel spacing is much larger than the Fourier frequency range over which the phase noise is to be controlled by the loop, otherwise, when locked to channels near the band centre, it is not possible to provide adequate suppression of the loop carrier frequency on the VCO output signal. As in the case of the sampling loop, then, the simple offset loop system is severely limited in channel capacity.

#### (c) Double Loop Schemes

To increase the number of channels obtainable from the sampling or offset loop schemes described above, a double loop approach may be employed. Here, a secondary loop is made to lock to offsets from a basic set of widely spaced channels generated by a primary loop. By choosing these secondary offsets as shown in Fig. 5, it may be arranged that the carrier frequency of the secondary loop is maintained sufficiently high compared with the loop bandwidth that the required spurious signal suppression is obtained for all channels selected.

This approach is highly successful, and in fact simplifies the design of the primary loop, since the filtering properties of the secondary loop reduce the need for a high degree of unused channel suppression in the primary channel set. The main disadvantages are complexity and degradation to phase noise and channel switching time, since the one loop must follow the other.

(d) Multiple Translation Loops

The basic offset scheme shown in Fig. 4 is an example of a system involving a single frequency translation within a phase locked loop. This may be extended to several translations within the loop, as demonstrated for example by Lewis [7,8] and Brown and Clarke [2]. The scheme devised by the latter two authors is outlined in Fig. 6, and uses directly synthesised frequencies in the HF/UHF range to generate 128 output frequencies across a 320 MHz bandwidth centred at around 3 GHz. As in the case of the simple offset loop, economies in the numbers of directly synthesised frequencies required are made by locking to the upper or lower sideband from each mixer.

(e) Schemes Based on Sidebands Generated by Mixing

Fig. 7 demonstrates the principle behind all schemes in this category. A microwave reference signal, usually at the agility band centre, is mixed with a directly synthesised signal at a lower frequency in a double-balanced mixer, resulting in suppression of the carrier and the generation of a pair of modulation sidebands. One of these sidebands is then selected by some means to yield the frequency agile output signal.

Selection of the wanted sideband may be carried out by an electronically tuneable or switchable filter, by image rejection in the mixer, or by a combination of the two.

The difficulty of designing an image rejection mixer with a sufficiently high IF bandwidth to achieve an agility range of 500 MHz may be overcome by arranging for the directly synthesised modulating signals to be generated in orthogonal pairs, rather than following the normal practice of including a 90° IF hybrid in the mixer. By carefully balancing the amplitude and phase of these pairs of signals, a very high degree of image rejection is possible. Further, since the microwave reference signal is of constant frequency, the use of feed-forward nulling is quite practical to achieve additional suppression of the LO signal.

Unfortunately, however, the conditions for total cancellation of the image do not in general result in complete suppression of the other, lower amplitude mixing products at harmonics of the IF offset from carrier, and in practice this results in residual spurious components around the -35 dBc level. Even when an image rejection mixer is used, then, some form of tuneable filter following the mixer is still required to obtain adequate far from carrier spurious signal suppression.

The double-mix arrangement shown in Fig. 8 and first described by Wadley [9] provides one solution to the tuneable filter problem. Cancellation of the VCO noise occurs, but this effect falls off as the offset from carrier increases, owing to the IF filter group delay, and eventually in fact, the noise is increased rather than cancelled. This results in a trade-off arising between channel spacing, far from carrier spurious rejection, and phase noise performance. Care must be taken with the choice of VCO frequency band to avoid the situation arising where the tuneable filter internal IF is close to any of the possible intermodulation products of the spectrum entering the first mixer, otherwise close to carrier spuri can occur. Alternatively, the VCO may be phase-locked, but this tends to abrogate the elegant simplicity of the Wadley scheme. Variants of the Wadley arrangement allowing the insertion of additional channels are employed in synthesisers described by Rauvola [10] and Mobbs [16].

An electronically tuneable filter may also be realised by employing a phase-locked loop as a tracking filter to select the desired modulation sideband from the initial mixer in the system. In this case much narrower bandwidths may be utilised than in the Wadley scheme, and the potential exists for increasing the number of channels further by introducing offsets into the loop. Such an approach has been described by McPherson, Walls and Brown [1], who have demonstrated two coherent radar synthesisers providing 17 and 33 channels from a single reference oscillator. The 17 channel scheme is reproduced in Fig. 9, and is attractive in the almost trivial nature of the direct synthesis required. Channels are derived in a similar manner to that shown for the double loop approach in Fig. 5, maintaining a high value of loop carrier frequency for all channels selected. The worst case far from carrier spurious signal level is quoted at -60 dBc for this scheme, but this could no doubt be very significantly improved by the use of an image rejection mixer rather than a double balanced mixer to apply the initial offset.

(f) Frequency Discriminator Schemes

Frequency locked loops based on transmission line type discriminators having periodic responses may be used as the basis of frequency agile microwave sources, rather than phase locked loops. Since in this case the discriminator becomes the fundamental reference controlling short term stability, it must be constructed using very high Q resonant lines.

An interesting and novel approach to a discriminator based system has been demonstrated by Haynes et al [11], employing a device described as a High Overtone Bulk Acoustic Resonator or HBAR. This is a piezo-electrically pumped acoustic equivalent of the optical Fabry-Perot resonator, and is used as the basis of a frequency discriminator with periodic responses. An extremely high value for the product of frequency and Q-factor has been reported for the HBAR device.

System Components and Engineering Considerations

Most of the frequency synthesis schemes which have been described in the preceding section depend critically on the performance of two key components, namely the reference oscillator and the microwave voltage controlled oscillator.

The current state of the art in bulk and surface wave piezo-electric crystal oscillators is indicated in Fig. 10. This figure has been compiled from various published results relating to high quality reference oscillators [12,13,14], all phase noise curves being scaled by ideal multiplication to 10 GHz for the purposes of direct comparison, although it is of course appreciated that the higher frequency oscillators would have some advantage in incurring less added noise in a real multiplication process.

The results suggest considerable competition between bulk wave and SAW resonator oscillators in the region between around 3 kHz and 20 kHz from carrier. Very close to carrier, bulk wave resonators seem to have the advantage, whereas in terms of noise floor SAW devices appear to consistently demonstrate superior performance. In most real applications, however, performance under vibration must also be taken into account, and here it does not yet seem possible to draw meaningful conclusions from published results. Reference [15] provides a useful recent review of a considerable volume of literature on the acceleration sensitivity of bulk wave quartz crystals. Much less information is available for surface wave devices.

Fig. 10 also shows the phase noise curves of three commercially available high quality X-band voltage controlled oscillators offering at least 500 MHz tuning range. The intersection of these curves with those of the reference oscillators indicates phase locked loop bandwidth requirements of around 5 to 10 MHz. To prevent degradation of the VCO noise performance,

great care must be taken in the design of any active circuitry connected to the tuning terminal. Summing of offset voltages for example is better accomplished by accessing the RF ground end of the tuning varactor, or injecting into the centre tap of a PSD transformer, than by resorting to the use of operational amplifiers. Similarly, active linearisers are best avoided by trying as far as possible to design the VCO with a linear tuning curve. To minimise noise appearing via the earthing system, the PSD and any subsequent dc gain should be located within the VCO package, preferably on the same substrate as the VCO itself.

Since the loop bandwidth of a phase locked system will generally be relatively high, perhaps of the order of 5 MHz, channel switching time will be largely dependent on the VCO post tuning drift and temperature drift characteristics, although the latter effect may be calibrated out.

The control of spurious signal levels demands that due attention must be paid to grounding, screening and filtering in all parts of the system. All control signals into the synthesiser, for example, should be optically isolated. Screening is particularly important in the case of direct synthesis, and here a conflict may occur between spurious signal performance and maintainability.

### Conclusion

Considerable progress has been made within the past four or five years in the design of high spectral purity microwave frequency agile sources for coherent radar applications. This has been driven by the ever increasing sophistication of modern radar signal processing techniques, made possible by the rapid advancement of digital technology. Performance of modern coherent radar systems has therefore become limited by those parts of the system which are analogue in nature, and in particular by any item degrading spectral purity.

In present generation systems based on TWT transmitters, particularly under conditions of high vibration, the microwave source and the transmitting tube represent the greatest threats to spectral purity. As solid state phased array systems appear, however, transmitter noise will be much reduced, and attention will become more and more focussed on the performance of the microwave frequency agile source.

In this paper the requirements expected of such a source have been examined, and several solutions to the synthesis problem reviewed. Results which have been published for some of these synthesis schemes are compared in Table 1. The figures listed in this table must however be used with caution in judging the merits of one scheme against another, since conditions in each case such as reference oscillator noise, standard of engineering, etc. will be different.

Looking to the future, we might expect developments in resonator devices and superconductivity to yield much lower noise reference sources than we have today. This in turn will result in synthesiser spurious signal specifications becoming even more demanding than at present. Vibration will continue to be a problem for designers of airborne systems, and any innovations in the resonator field offering lower acceleration sensitivity will be received with great interest.

The impact of GaAs digital technology will undoubtedly result in changes to present day techniques, and in this context use of the direct digital frequency synthesiser (DDFS) will become much more common place, and may open up new methods. The ability of the DDFS to generate non-sinusoidal waveforms, for example, could enable near perfect spurious suppression to be

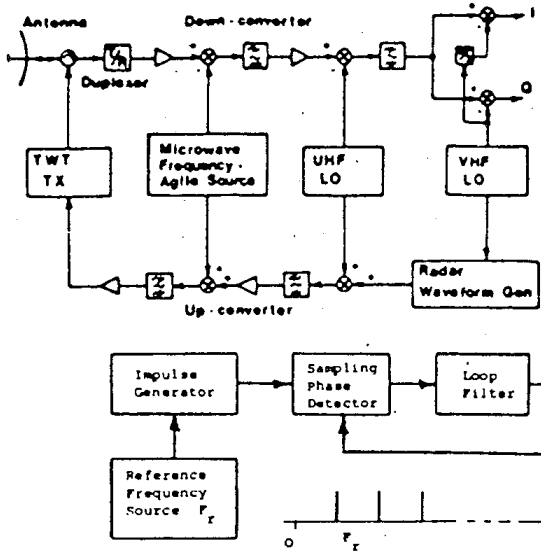
achieved in image rejection frequency translation by pre-distortion of the modulating IF signal to compensate for mixer imperfections.

The development of systems with high resolution or imaging capabilities are likely to generate a need for greater numbers of frequency channels, and perhaps wider agility bandwidths. In many cases however the latter parameter is unlikely to increase substantially. Although the solid state power amplifiers of phased array radars will have broader band capability than travelling wave tubes, the need for good antenna performance will impose a bandwidth restriction.

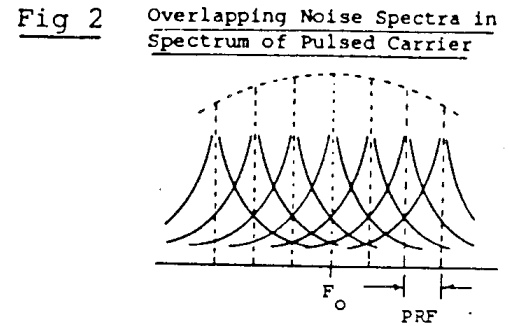
Finally, an expected growth in millimetre wave systems will create an increasing need for precision synthesis techniques in this area.

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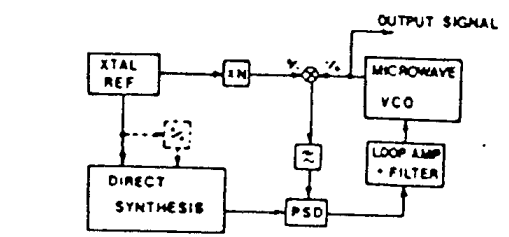


**Fig 1**  
Typical Microwave/RP System Architecture in a Coherent Radar System

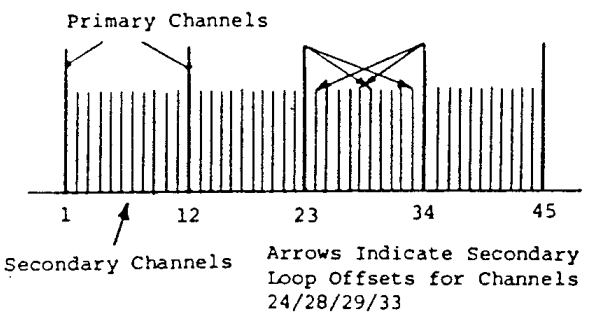


**Fig 2**  
Overlapping Noise Spectra in Spectrum of Pulsed Carrier

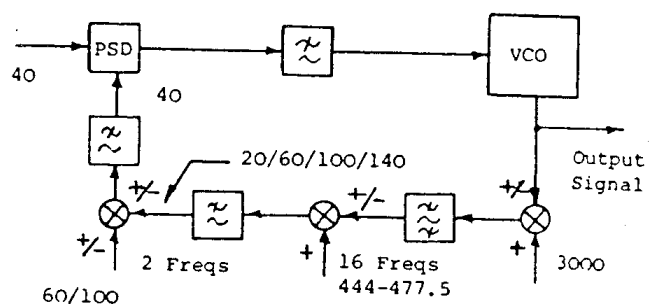
**Fig 3** Basic Sampling Loop



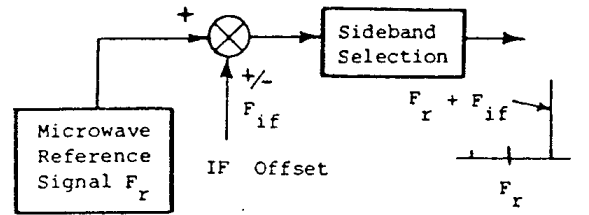
**Fig 4** Simple Synthesis Scheme with Directly Synthesised Offsets



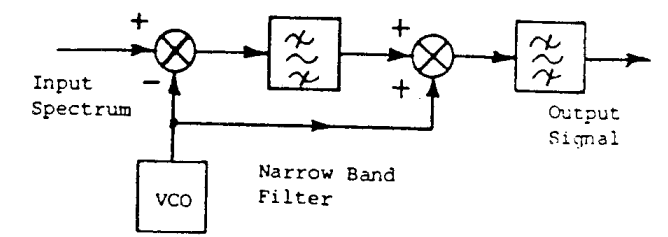
**Fig 5** Typical Channel Derivation for Double Loop Synthesis



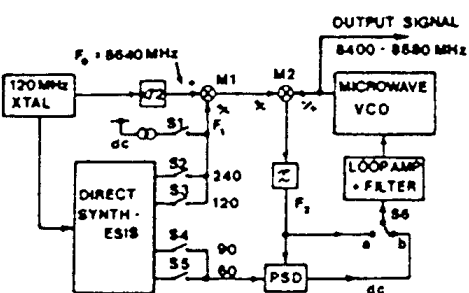
**Fig 6** Multiple Translation Loop (from ref 2)



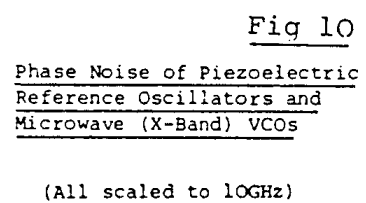
**Fig 7** Synthesis Based on Sidebands Generated by Mixing



**Fig 8** Wadley Tuneable Filter Arrangement



**Fig 9** 17 Channel Synthesiser (from ref 1)



**Fig 10**  
Phase Noise of Piezoelectric Reference Oscillators and Microwave (X-Band) VCOs  
 (All scaled to LOGHz)

**Table 1**

Performance of Some Microwave Frequency Agile Sources

\* Scaled to LOGHz  
 \*\* From Single Ref Osc

Scheme	No of Channels	Centre Freq (GHz)	Agility Band (MHz)	* Phase Noise (dBc/Hz)			Max Spur Level (dBc)	Switch Time (µsec)
				2K	20k	100K		
Sampling Loop(ref 4)	3	X-Band	—	-92	-110	-107	-77	250
Mult Trans Loop(ref 2)	128	3	338	-106	-115	-115	-100	8
Mixing + Offset Loop(ref 1)	33	8,64	500	-117	-127	-130	-60	25