SOLID-STATE LINEAR FM/CW RADAR SYSTEMS - THEIR PROMISE AND THEIR PROBLEMS

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Abstract

The problems associated with linear FM/CW solid state radars are examined in detail, particularly sweep non-linearities. Design information is given which predicts the range sidelobe levels of such radars and methods of reducing these to acceptable levels are discussed.

Introduction

The advent of solid-state microwave power sources has brought with it a considerable interest in short-range radar systems. The early simple intruder alarm type radars are now being followed by longer range, more sophisticated equipments as higher transmitter powers become available. One of the most significant features of these new sources is that the CW powers available are at least 20 times greater than the mean powers obtainable from pulsed devices (Fig. 1). This difference is fundamental and is due to either voltage breakdown or thermal limitations in the pulsed sources. Therefore, when maximum mean transmitter power is required there is a clear advantage in choosing a CW source which, of course, has to be modulated if range information is required.

Many system designers have been deterred from implementing a linear FM/CW system in the past because of transmitter sweep linearity problems and the fact that twin antennas are usually required. However, because this system can offer a maximum transmitter power capability, there is a good case for re-examining the technical problems. This paper sets out to do just this and show that once the problems are understood they are not insuperable using modern technology.

In a linear FM/CW radar, target range is deduced by noting the difference in frequency between transmitted and received signals (Fig. 2). For the more usual multiple target case many difference frequencies will be present in the receiver simultaneously and these have to be resolved by some form of spectrum analyser. These difference signals are relatively low frequency, usually less than 20 MHz, and are entirely independent of the system range resolution. This means that the receiver, after the first mixer, contains no novel or complex technology.

The principal technical problem, associated with this type of system, is that of transmitter sweep linearity as any non-linearities of the FM waveform produce spurious range sidelobes in the spectrum analysis. Therefore, in this paper we shall present design curves which can be used to determine the linearity requirements for most practical equipments and discuss how these requirements can be implemented.

Ideal System

Before introducing practical imperfections it is essential to know the resolution and range sidelobe levels of an ideal system. As can be seen on Fig. 2, the envelope of the frequency spectrum of a train of rectangular bursts of the difference frequency signal (f_d) is the well-known sinx/x function. This means that the first "range sidelobes" are only some 13 dB down on the wanted signal. This is usually inadequate if targets of unequal amplitude are to be resolved and some form of amplitude weighting of the signal bursts (such as Hamming) has to be used to reduce these to an acceptable level. This weighting can also be used to compensate for the effects of transmitter amplitude modulation. It will be assumed in the following discussion that this precaution has been taken. One unfortunate consequence of using amplitude weighting is that the main lobe is broadened which reduces the theoretical system resolution. For Hamming weighting, it can be shown that the target resolution (δR) is approximately given by the velocity of propagation divided by the swept bandwidth (B).

Non-Linear Frequency Sweep

In any practical system, the frequency sweep can never be made quite linear. This deviation from linear, as a function of time, is usually irregular but as it is periodic it can be expressed as a Fourier series of sinusoidal components of frequency m/T, where T is the sweep time and m is an integer (see Fig. 6). The amplitude of each component (k_m) is the peak frequency deviation from linear. An example of this is shown on Fig. 3 for the case of a single sinewave with m equal to 4. The phase of the modulation relative to the start of the sweep is given by ϕ_4.

The result of this unwanted modulation is to produce spurious frequency components in the received spectrum. For instance, if k_4 is made small and ϕ_4 zero then a single pair of spurious lines appear equally spaced about the wanted difference frequency (f_d) at frequencies 4/T relative to it. If ϕ_4 is made $\pi/2$ then the amplitude and number of spurious lines appearing changes somewhat but this is a second order effect. The amplitude of the lines depends primarily on the values of the peak frequency deviation (Fig. 4 shows this for k_4) and on the time delay between transmit and receive (t_d). In fact, it can be shown that the effective peak deviation (y_m), as seen by the receiver, is given by:

$$y_m = \frac{k_m . T}{m\pi} \sin\left(\frac{m\pi}{T} . t_d\right) \quad (1)$$

If $\frac{m\pi t_d}{T}$ is very small then the equation reduces to:

$$y_m = k_m . t_d \quad (2)$$

This is a very important result because it means that, providing this approximation is valid, the magnitude of the interference due to a sinewave frequency modulation of the transmitter is dependent only upon the maximum deviation at the transmitter and the target range. Note, particularly, that it is independent of other system parameters such as band swept, sweep time and even of the frequency of the unwanted modulation. This last parameter does, of course, determine in which frequency (ie: range) cell the interference appears but it does not influence its magnitude. It was found that there was less than a 1% error in the results using Equation (2) rather than (1) providing that:

$$\frac{m\pi t_d}{T} \leqslant \frac{\pi}{10} \qquad (3)$$

For the vast majority of practical cases, this condition can easily be met for the simple reason that the time to the furthest target of interest is nearly always made a small fraction of the sweep time to avoid the problems of second-time round returns.

Using results similar to the examples shown on Fig. 4, it is now possible to draw a set of curves which relate the transmitter non-linearities to the maximum sidelobe levels generated for any given target range. These are given on Fig. 5.

This, of course, is not the whole story because these design curves relate to a single sinewave modulating the linear frequency sweep and, as shown on Fig. 6, a practical modulation consists of an infinite series of such sinewaves of various amplitudes. However, it is usual for one or two components to be significantly greater than the remainder and on all the practical examples examined it has been found that it is reasonable to allow for the multiple frequency nature of the waveform by decreasing the allowable transmitter deviation by a factor of 2.

As an example, consider the case of a radar which is required to have a resolution of 1 metre at a maximum range of 100 metres. The range (or interference) sidelobes are not to exceed -20 dB on the peak signal from the target. From Fig. 5, and remembering to divide by 2, the maximum allowable deviation from linear must not exceed 20 kHz. The desired resolution requires at least a bandwidth of 300 MHz. Hence the required tolerance on the frequency sweep linearity of the transmitter may be expressed as ±0.007% where this should be taken to mean the peak deviation irrespective of waveform. Reference must be made to Equation (3) to ensure that the results are valid. For this we need to know the sweep time which is very unlikely to be less than 100 µS. Therefore, the results obtained start to contain errors only if modulation components of 20 cycles or greater, having significant amplitude, are present. This is very unlikely.

Practical Limitations on Transmitter Linearity

There are only two satisfactory methods, at present, of frequency tuning a solid-state microwave oscillator - varactor tuning and YIG tuning. YIG tuned oscillators have a wide linear tuning range and a high Q; up to ±0.2% linearity over octave bandwidths have been achieved. However, they are limited in speed of modulation to about 1,000 sweeps/sec though this can be increased to about 30 kHz if the sweep bandwidth is limited to, say, 20 MHz by using a low inductance tuning coil. Note that these figures apply to sinusoidal sweeping of the oscillator; for a sawtooth waveform the rates are about an order of magnitude less. The varactor tuned oscillator, on the other hand, allows sinusoidal modulation rates up to several MHz and sawtooth sweeps up to at least 100 kHz but the frequency is not linear with tuning voltage because of the varactor law. Some form of lineariser is, therefore, essential for this type of tuning, even to achieve 1% linearity and it is also necessary with the YIG oscillator if better than 0.1% is required.

The first method is to apply an open loop correction to the tuning voltage (or current) waveform. This technique is particularly applicable to the varactor where the addition of the conjugate waveform to the varactor tuning law should, theoretically, bring it up to at least the performance of the YIG. This conjugate waveform consists of an infinite series of components of which perhaps the first ten or twenty are significant. As an example, the addition of only four terms in correct amplitude and phase to compensate for the varactor law has been shown experimentally to improve the linearity to better than 0.1% over a 300 MHz sweep. If, however, much better linearity is required, then this technique is inadequate and some form of closed loop compensation must be used.

Closed loop feedback systems can be divided into two classes: those which operate in real-time on the sweep as it is generated and those which use the errors generated on one sweep to correct subsequent sweeps. Both classes depend on being able to measure the error in linearity so that it can be used as the feedback variable and this is usually done by using a fixed delay (t_d) in the transmitter circuit to generate a difference frequency whose phase is compared in a phase sensitive detector with a reference. The minimum value of t_d is determined by the measurement sensitivity of the phase sensitive detector (Θmin) and is given by:

$$t_d \doteq \frac{1}{2\pi} \cdot \frac{\Theta_{min}}{\Delta f} \qquad (4)$$

An upper limit is imposed on t_d if high frequency components of significant amplitude are present. Linearities of at least ±0.01% are achievable with closed loop systems but this depends on the sweep time T because the effectiveness of the feedback control decreases with T. Increasing the loop gain can alleviate this providing stability can be maintained. If T is 10 µsecs, for instance, then real time feedback linearisation is not very useful. The choice of sweep time T is a compromise between several factors. The first is that T must be long enough so that second time round echoes are not visible. This suggests as slow a tuning rate as possible and this is also desirable for practical reasons. However, a very slow sweep implies a narrow bandwidth receiver and this may be undesirable because it must be wide enough to accept any target modulation such as glint; hence T must usually be much less than 1 second. A slow sweep also means that the difference frequencies generated are low, particularly for minimum range targets, and have to compete with the 1/f noise of the mixer, which is very significant close to the carrier. Experiments have shown that the difference frequencies should not be less than 10 kHz to avoid this problem.

The second class of control loop stores the error signal in a digital form, which is continually updated and is used to correct subsequent sweep cycles. This method does not suffer from the conflicting requirements of "real time" loops and is of considerable interest because it has great flexibility and potential, as well as promising to achieve very linear fast sweeps.

Conclusions

The transmitter linearity required is the chief problem of high resolution FM/CW radars but, with the design information presented here, it is now possible to implement the sweep controls necessary to realise practical systems. Therefore, the potential of being able to double the radar range obtainable by any solid state pulse system is soon likely to be achieved.

Acknowledgements

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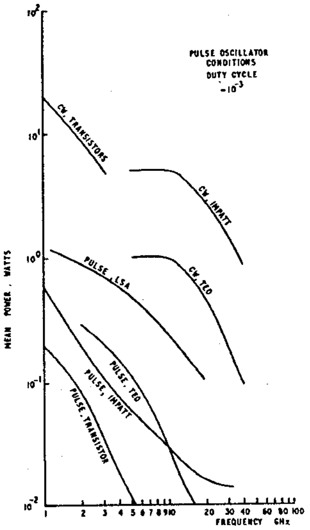

FIG.1.
COMPARISON OF MEAN POWERS GENERATED
BY PULSED AND CW SOLID STATE OSCILLATORS.

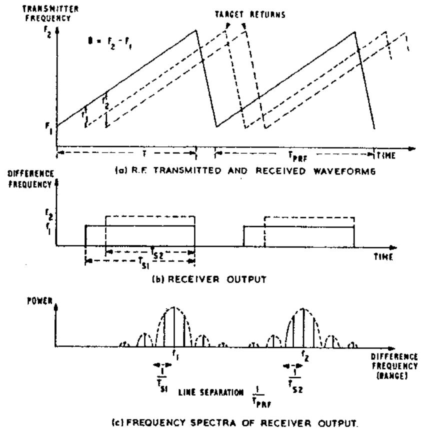

FIG.2.
BASIC FM/CW RADAR WAVEFORMS SHOWING TWO WIDELY SEPARATED
TARGETS.

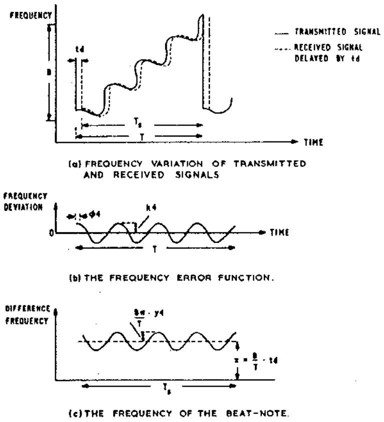

FIG. 3
THE EFFECT OF SINUSOIDAL FREQUENCY
MODULATION OF THE TRANSMITTER SWEEP.

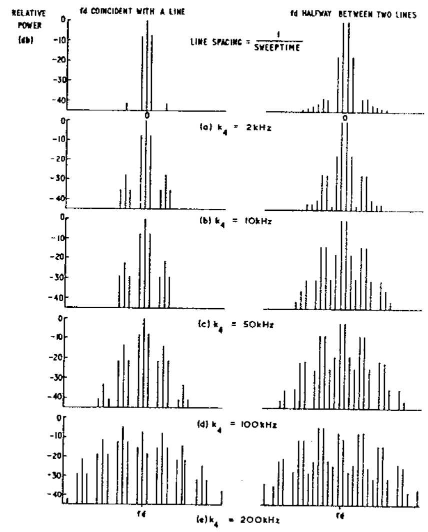

FIG. 4
VARIATION OF INTERFERENCE AMPLITUDE WITH DEVIATION
FROM A LINEAR SWEEP.

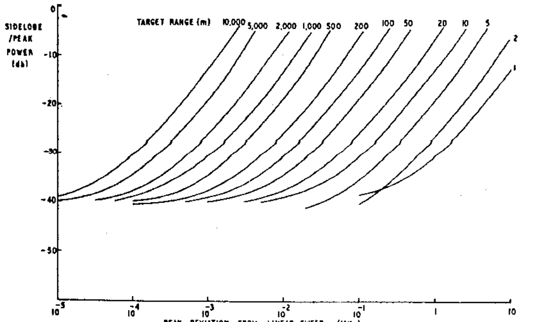

FIG. 5
VARIATION OF SPURIOUS SIDELOBE LEVEL WITH TARGET RANGE AND PEAK TRANSMITTER FREQUENCY DEVIATION.

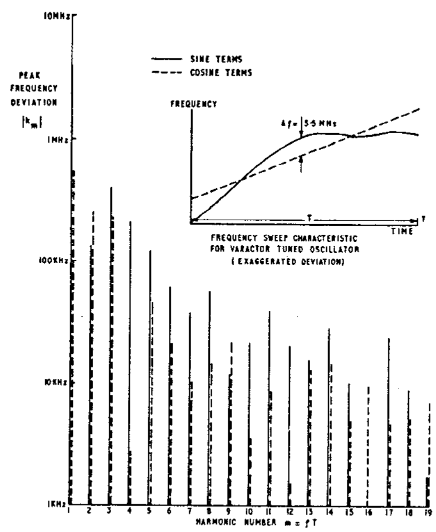

FIG. 6.
FREQUENCY DEVIATION SPECTRUM FOR VARACTOR
TUNED OSCILLATOR.