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MODULATION WAVEFORMS FOR CONTINUOUS WAVE RADAR,(U)
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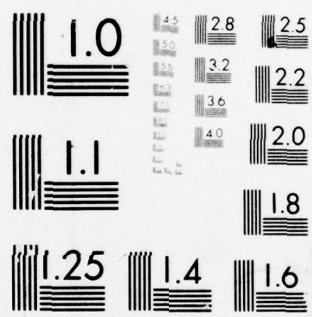

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MODULATION WAVEFORMS FOR CONTINUOUS WAVE RADAR

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Fort Monmouth, New Jersey

1974

6p.

ABSTRACT

The range of a radar is a function of average transmitter power and is independent of the modulation waveform. It does not matter whether the transmitter operates in a pulsed or CW mode. In many applications, an average power of only a few watts is adequate. In these cases, CW operation can have a number of advantages. Solid state power sources can readily be utilized and a single RF source can serve as both transmitter and local oscillator. Another advantage is waveform flexibility. In a CW radar the waveform can be designed to perform a variety of functions, e.g., detect fixed and moving targets, determine sense of radial velocity and communicate with a target over a fade-free channel.

This paper describes waveforms and design techniques that can be used to achieve the above mentioned functions with a CW radar.

SIMPLE CW RADAR

A simple CW radar utilizing a common antenna for transmit and receive is shown in Figure 1. (For simplicity, in Figure 1 and all subsequent block diagrams, no amplifiers are shown.) The RF GENERATOR and RF DEMODULATOR are coupled to the RF ANTENNA by means of an RF CIRCULATOR. Local oscillator power for the RF DEMODULATOR is obtained by leakage through the RF CIRCULATOR or directly from the RF GENERATOR. The useful output of the RF DEMODULATOR will be doppler frequency signals from moving targets within the radar beam.

A radar as shown in Figure 1 is not practical for most applications. Transmitter and mixer diode noise close to the RF carrier will fall within the doppler frequency band, resulting in poor receiver sensitivity. Since it is not possible to employ conventional sensitivity time control in the receiver, strong signals from nearby targets will mask weak returns from distant targets. Also, it is not possible to determine the range of targets. Finally, fixed target detection is impractical, since highly stable dc amplifiers would be required.

In the next several sections we will show how the above limitations can be overcome by frequency and phase modulation of the transmitter. First we will discuss solution of problem of masking by nearby targets.

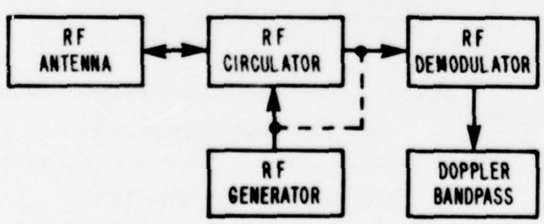


FIG. 1 : SIMPLE CW RADAR

SINE WAVE FM

Target output signal strength can be made almost independent of range (over certain limits) by frequency modulating the transmitter with a sine wave and processing received signals, as shown in Figure 2a. Again, the RF GENERATOR is coupled to the RF ANTENNA via the RF CIRCULATOR. Some of the transmitted signal leaks through the RF CIRCULATOR into the RF DEMODULATOR and provides a reference for the detection of returned echoes. The output of the RF DEMODULATOR is given by (Refs. 1,2).

$$E_{fm} = \sqrt{A}\,\left(\frac{T}{\tau}\right)^2 \left[E_e(t)\cos(2\pi f_o \tau) - E_o(t)\sin(2\pi f_o \tau)\right] = E_I - E_Q \quad (1)$$

where A = constant representing fixed factors in the radar range equation

$$T = \frac{1}{f_m} = \text{modulation freq period} \quad (2)$$

τ = round trip delay time of target signal

$$E_e(t) = J_o(m_f) - 2\left[J_2(m_f)\cos(2\,\theta(t)) + \ldots \quad (3)$$

$$E_o(t) = 2\left[J_1(m_f)\cos(\theta(t)) - J_3(m_f)\cos(3\,\theta(t)) + \ldots \quad (4)$$

J_n = Bessel function of first kind, order n

$$m_f = \frac{2\,\Delta f}{f_m}\sin(\pi\,\tau\,f_m) \quad (5)$$

Δf = peak frequency deviation of transmitter

$$\theta(t) = 2\pi f_m t - \pi f_m \tau \qquad (6)$$

f_o = center frequency of transmitter

E_I = In-phase component of demodulated RF

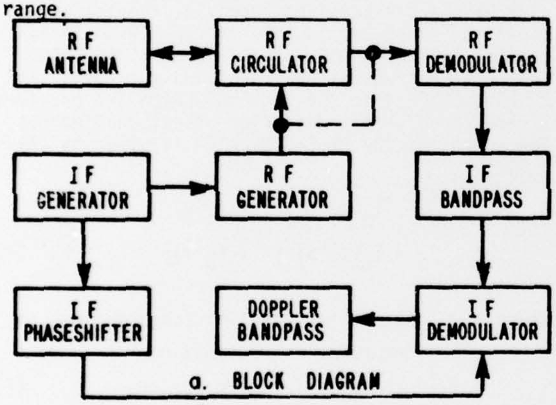
E_Q = Quadrature-phase component of demodulated RF

The IF BANDPASS is tuned to f_m, thus eliminating much of the low frequency system noise.

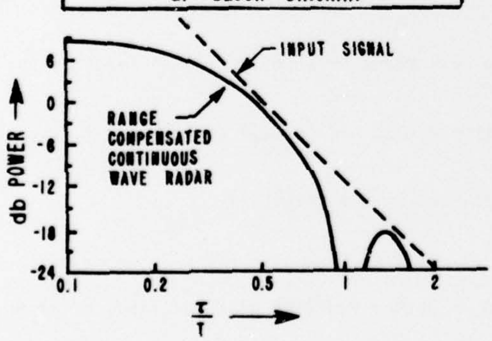
It can be shown (Ref. 1) that the output of the IF DEMODULATOR is given by

$$E_d = \sqrt{A}\left[\left(\frac{T}{\tau}\right)^2 \sin\left(\pi\frac{\tau}{T}\right) J_1\left(\frac{2\,\Delta f}{f_m}\sin\left(\pi\frac{\tau}{T}\right)\right)\right] \sin(2\pi f_0 \tau) \qquad (7)$$

The term in brackets represents the variation of target signal output as a function of delay time. A plot of the square (proportional to power) of this function vs τ / T is shown in Figure 2b for the case where $2\,\Delta f / f_m$ is equal to 1.8. The minimum processing loss (about 2 dB) occurs when τ / T is 0.5. A design criterion is then that the maximum target delay time be equal to $\frac{1}{2}T$. The signal output varies by about 10 dB as the target moves from maximum to zero range. Also shown in Figure 2b is the input signal strength, which decreases at the rate of 12 dB per octave of range.



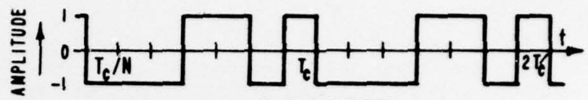
a. BLOCK DIAGRAM



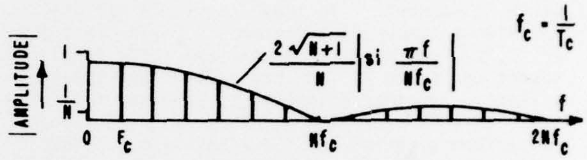
b. POWER OUTPUT VERSUS RANGE DELAY TIME

FIG. 2: CW RADAR WITH RANGE COMPENSATION

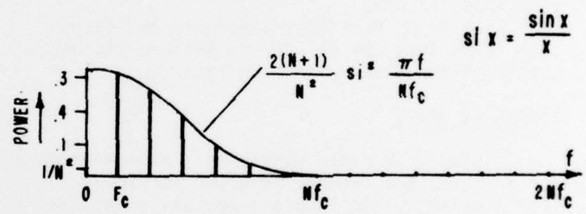
Two factors contribute to the signal compression. First, for small arguments the first order Bessel function is very nearly proportional to τ, thereby giving a 6 dB reduction in signal each time the range is halved. Another 6 dB reduction is obtained from the $\sin(\pi\,\tau / T)$ term. This term comes about through the change in phase of the IF signal with range.



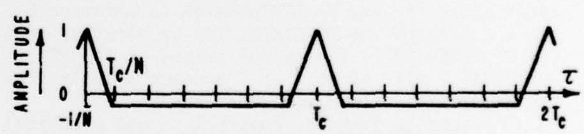
a. TIME FUNCTION



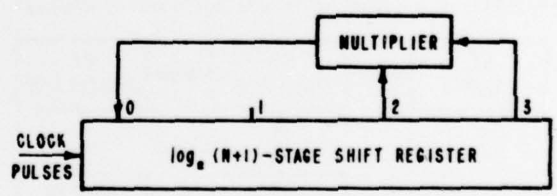
b. AMPLITUDE SPECTRUM



c. POWER SPECTRUM



d. AUTO-CORRELATION FUNCTION



e. N ELEMENT BINARY CODE GENERATOR

FIG. 3: MAXIMUM LENGTH PSEUDO RANDOM CODE

RANGE RESOLUTION

In order to obtain range resolution in a CW radar, it is necessary to modulate the transmission. A suitable modulation waveform is a maximum length binary pseudo-random code (Refs. 3,4). A seven-element code is shown in Figure 3. Since the pseudo-random code is quantized in both amplitude and time, it can be conveniently generated by a shift register with feedback. The feedback is accomplished through binary multiplication of the outputs from two of the shift register stages, using the product as the input to the first stage. The operation of a multiplier is such that its output is positive if the inputs are of like polarity, and negative if the inputs are of opposite polarity. The code length N is 2^n-1 elements, where n is the number of stages in the shift register.

A three-stage shift register with feedback from the second and third stage is shown in Figure 3e. With every clock pulse the information in each stage of the SHIFT REGISTER is transferred to the following stage. If an output is taken from one stage, a code will be generated which repeats every seven clock pulses.

From Figure 3e, it is evident that the output from the second stage of the SHIFT REGISTER is delayed by one clock period (time between successive clock pulses) from the output of the first stage. Hence, any desired amount of delay can be obtained by cascading additional stages. A delay can also be obtained by using a second shift register and delaying its clock pulses with respect to those of the first.

The code has three interesting features. If we represent the code waveform by $u(t)$, then when τ_1, τ_2, and τ_3 are multiples of a clock period,

$$u(t) \cdot u(t) = 1 \tag{8}$$

$$u(t-\tau_1)\, u(t-\tau_2) = u(t-\tau_3) \tag{9}$$

$$\overline{u(t)} = \frac{1}{N} \tag{10}$$

Equation 8 says that a code multiplied by itself yields a constant value of unity, while Equation 9 says that the product of two shifted codes is again a code, but with a new shift. Equation 10 says that the mean value of the code is 1/N. These three features give rise to an autocorrelation function, as indicated in Figure 3d, with a sidelobe level of 1/N. For example, if N is 1023 the theoretical peak to sidelobe ratio is over 60 dB. In practice, it is difficult to meet the component tolerances required to obtain more than 50 dB.

In some applications a side lobe level of 50 dB is not adequate. A technique for increasing the rejection of off-range targets is discussed in the following section. This technique is applicable when the highest doppler frequency is less than one tenth of the code repetition frequency.

COMBINATION OF PSEUDO-RANDOM CODE MODULATION AND SINE WAVE FM

The sine wave FM can be combined with the pseudo-random code modulation if certain precautions are taken. A block diagram of the combined system is shown in Figure 4a. The RF GENERATOR is sine wave frequency modulated. The output of the RF GENERATOR is applied to the RF MODULATOR, where it is balanced modulated by the CODE GENERATOR. The RF MODULATOR changes the phase of the transmitted signal by 180° each time the code changes state. The RF MODULATOR is coupled to the RF ANTENNA via the RF CIRCULATOR. Again, transmitter leakage through the RF CIRCULATOR into the RF DEMODULATOR is the product of transmitted code times the returned code modulated by an FM waveform. The modulation impacted by the mixing of the transmitted and received signals in the RF DEMODULATOR is accounted for by multiplying the delayed code by the transmitted code in the CODE MULTIPLIER of Figure 4a. The output of the CODE MULTIPLIER is applied to the CODE DEMODULATOR. If the CODE DELAY is τ', the output of the CODE DEMODULATOR is then

$$E_s(t) = (E_I - E_Q)\, u(t-\tau)\, u(t-\tau') \tag{11}$$

Since the IF BANDPASS will pass only frequencies close to f_m, it is necessary to examine the spectrum of $E_s(t)$. The spectrum of $E_s(t)$ is the convolution of the code spectrum with harmonics of f_m. In order to maintain the 1/N peak-to-sidelobe

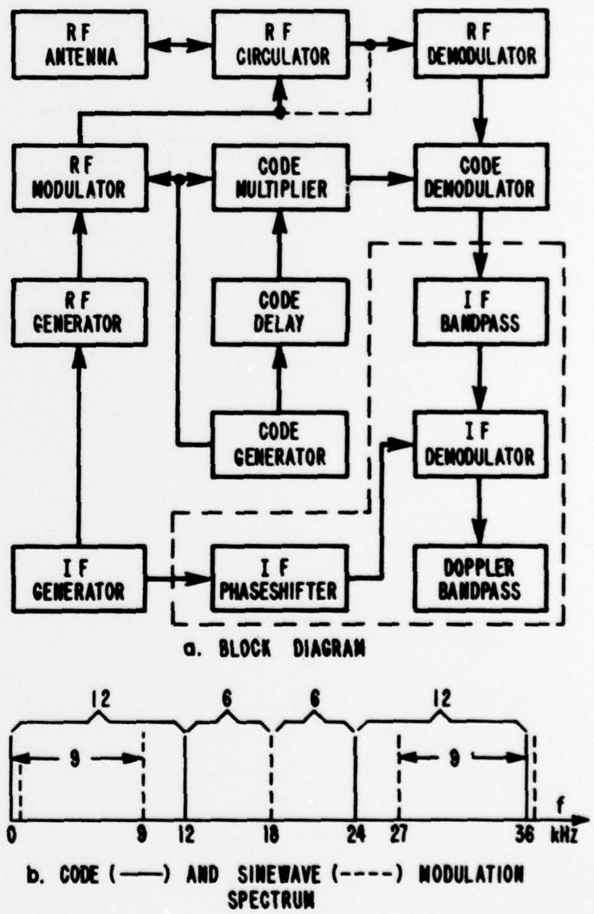


FIG. 4 : CW RADAR WITH RANGE RESOLUTION

as the target doppler frequency is below 100 cps, the speech intelligibility will not be appreciably affected.

If "a" and "b" are not equal, then the signal amplitude will change with target range, but will never be less than the smaller of "a" or "b".

CONCLUSIONS

The techniques we have discussed illustrate the waveform design flexibility available in a CW radar. The application and extrapolation of these techniques can improve the performance of many types of radars; e.g., surveillance, collision avoidance, marine, intrusion detection and police.

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2. W. Fishbein, O. Rittenbach, "Multifunction Radar Waveforms," EASCON '69 Record.

3. W. Fishbein, O. Rittenbach, "Correlation Radar Using Pseudo-Random Modulation," 1961 IRE International Convention Record.

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ratio of the pseudo-random code, it is necessary that the f_m component in the spectrum of $E_s(t)$ arise only from the product of the J_1 term of E_Q and the dc component of the code spectrum. As can be seen from Figure 3b, if harmonics of the code repetition frequency give rise to an f_m component in $E_s(t)$, the sidelobe level of the autocorrelation function will be down only by approximately $1 / \sqrt{N}$ instead of $1 / N$.

The above mentioned requirement on the spectrum of $E_s(t)$ can be satisfied by taking advantage of the rapid decrease in amplitude of the higher order Bessel functions. For all practical purposes the spectrum due to the frequency modulation (the spectrum of $E_I - E_Q$) is finite. If we keep m_f below 1.5, then we need consider only frequencies up to 3 f_m.

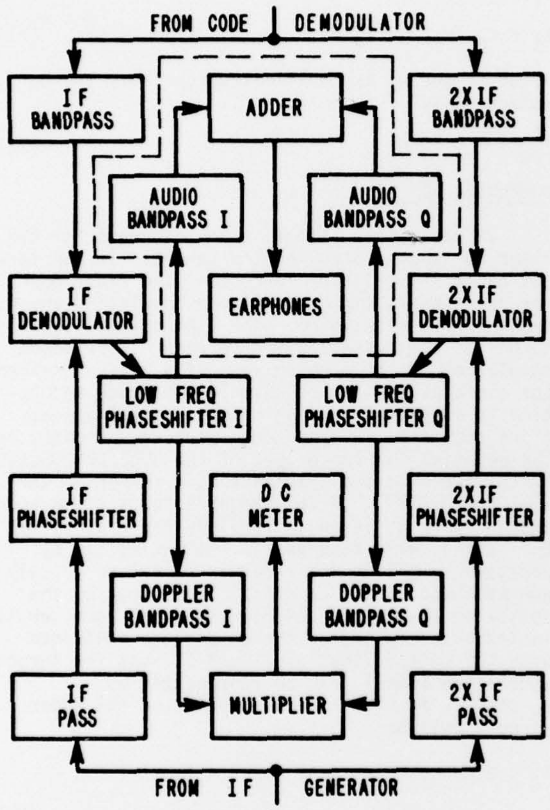

FIG. 5 : SINGLE SIDE BAND PROCESSOR

The procedure for the selection of f_m and the code repetition frequency, f_c, is best illustrated by an example. Consider the case of a radar which is required to detect targets out to a range of 10 km. A suitable f_c is 12 kHz. If f_m is 9 kHz, then an undesired 9 kHz signal can arise from a beat between the 3rd harmonic of f_c (36 kHz) and the 3rd harmonic of f_m (27 kHz). The relationship of the various frequencies is shown in Figure 4b.

The voltage ratio of the desired to undesired signals is given by the following expression:

$$R_{13} = J_1\ (m_f)\ /\ (\sqrt{N}\ J_3(m_f)) \tag{12}$$

If m_f is limited to 1.5 at maximum range, then R_{13} will be at least 35 dB even when the code length is as short as 63. As the target range decreases m_f will decrease and the J_3 component will drop off rapidly. The drop off will be more rapid than the increase in returned power, due to decreasing range. Thus the rejection ratio of off-range targets will be at least 35 dB at all ranges. A 9 kHz signal can also arise from the 5th harmonic (45 kHz) of f_m, however, its amplitude will be negligible.

The choise of 9 kHz and 12 kHz limits the available doppler band to about 1000 cps. The limitation on the doppler band is due to the fact that close-in targets will produce strong code lines in the output of the CODE DEMODULATOR. If target has a doppler frequency of 1000 cps, the code will have a component of 11 kHz. For a 63 element code, this component will be down by only 18 dB from the entire signal. When the 11 kHz component mixes with the 9 kHz reference in the IF DEMODULATOR, the beat frequency will be 2 kHz. To maintain an 80 dB peak to sidelobe ratio, the 2 kHz component must be attenuated 62 dB by the DOPPLER BANDPASS. If the cut-off frequency of DOPPLER BANDPASS is 1000 Hz, the required attenuation can be obtained with a small filter. If desired, some of the attenuation of the 11 kHz signal can be effected in the IF BANDPASS.

In order to avoid any spurious signals, it is essential that all beat frequencies between oscillators in the system be a multiple of 3 kHz. For example, the power supply and the oscilloscope sweeps should operate on harmonics of 3 kHz. To lock f_m and f_c, they should be generated from the same oscillator.

FIXED TARGET DETECTION

In order for fixed targets to be detected by the radar of Figure 4 they must be modulated by a frequency which falls within the doppler passband. Fixed targets can be modulated by frequency modulating the RF GENERATOR with a second sine wave. If the frequency of this sine wave is f_F and the peak transmitter frequency deviation is Δf_F, it can be shown that the output of the CODE DEMODULATOR, $E_s(t)$, will be modified such that

$$\begin{matrix}\cos\\ \sin\end{matrix}\left[2\ \pi\ f_o\ \tau\right]$$

is replaced by

$$\begin{matrix}\cos\\ \sin\end{matrix}\left[2\ \pi\ f_o\ \tau + \frac{2\ \Delta f}{f_m} \sin(\pi\ f_F\ \tau)\ \sin(2\ \pi\ f_F\ \tau - \pi\ f_F\ \tau\right] \tag{13}$$

The above expression represents an FM waveform with approximately equal spectral components, out to $2 \pi \Delta f_F \tau f_F$. For example, if the doppler frequency passband of interest is between 30 and 1000 Hz and the maximum range is 10 km, a suitable set of parameters if f_F = 30 Hz, Δf_F = 75 kHz. Then the highest frequency in the FM waveform will be 900 Hz. Thus, fixed targets will be modulated at frequencies between 30 and 900 Hz and will pass through the doppler filters.

SENSE OF RADIAL VELOCITY

In many operational situations it is useful to know if the target is approaching or receding. The well known doppler principle states that the frequency of the reflected signal is higher than the transmitted signal when the target is approaching and lower when it is receding. In order to distinguish approaching from receding targets, it is necessary to have a radar receiver which can tell whether the frequency of the reflected signal has increased or decreased.

Sense of radial velocity can be determined by processing quadrature components of the RF signal. As shown by the equation for the output of the RF DEMODULATOR, $E_s(t)$, quadrature components of the RF signal, $\cos(2 \pi f_o \tau)$ and $\sin(2 \pi f_o \tau)$, are available. The sine component appears at the output of the IF DEMODULATOR (Figure 4a). The cosine component can be obtained by processing the second harmonic of the FM waveform, i.e., the J_2 component. The second harmonic is processed in essentially the same manner as the first harmonic (see Figure 5). The only difference besides using a bandpass centered at $2 f_m$ is the setting of the IF Phase Shifter. For small arguments, the second order Bessel function decreases by 12 dB each time the argument is halved. Thus, the amplitude of the J_2 component of the returned signal will be nearly constant with range. The phase of the J_2 component varies by 180^o as the target delay changes from zero to $\frac{1}{2}T$. It is then necessary to couple the IF PHASE SHIFTER to the range delay control in order to maximize the signal.

A block diagram of the sense-of-radial-velocity system is given in Figure 5. The components in Figure 5 replace those within the dotted lines of Figure 4a. The dependence of meter deflection upon target direction can be demonstrated by the following analysis:

Let the output of the IF channel be represented by

$$E_1 = K_1 \sin(2 \pi f_o \tau) \tag{14}$$

and the output of the 2 IF channel by

$$E_2 = K_2 \cos(2 \pi f_o \tau) \tag{15}$$

If the target moves, τ will vary as a function of time. The outputs will then be

$$E_1 = K_1 \sin(\phi + 2 \pi f_d t) \tag{16}$$

$$E_2 = K_2 \cos(\phi + 2 \pi f_d t) \tag{17}$$

where f_d is the target doppler frequency and ϕ is a constant phase angle. No generality is lost by dopping out ϕ in both terms. Then:

$$E_1 = K_1 \sin(2 \pi f_d t) \tag{18}$$

$$E_2 = K_2 \cos(2 \pi f_d t) \tag{19}$$

f_d will be positive if the target is receding and negative if the target is approaching.

As shown in Figure 5, each signal is passed through a low frequency phase shifter. The design of the phase shifters is such that, over the band of frequencies of interest, the differential phase shift between the channels will be 90^o. The outputs of the phase shifters will be:

$$E_1 = \pm K_1 \cos(2 \pi f_d t) \tag{20}$$

$$E_2 = K_2 \cos(2 \pi f_d t) \tag{21}$$

After E_1 and E_2 are multiplied together, the DC component is

$$E_{DC} = \pm \tfrac{1}{2} K_1 K_2 \tag{22}$$

COMMUNICATION LINK

It may be operationally advantageous for the radar to communicate with the target. If the target has a crystal video detector, the radar operator can contact him by amplitude modulating the CW transmission with voice. The simplest way for the target to communicate with the radar is by voice modulation of the echoing area of a passive reflector aimed at the radar. Simple amplitude modulation is subject to fading unless special precautions are taken in the radar receiver. Looking at the equation for the output of the CODE DEMODULATOR $E_s(t)$, the AM signal, appears as a change in the amplitude of $\sqrt{A}$. If the target is at a range such that $2 \pi f_o \tau$ is an integral multiple of π, the E_Q signal will disappear. The fading can be overcome by making the receiver sensitive to only one sideband of the AM signal. As shown by the following analysis, a single side band receiver is instrumented by adding the components contained with the dotted lines of Figure 5. Let the target modulation signal, $\sqrt{A}$, be represented by $E_v \cos(2 \pi f_v t)$. Then the output of the Adder circuit will be:

$$E_A = a E_v \cos(2 \pi f_v t) \cos(2 \pi f_o \tau) + b E_v \sin(2 \pi f_v t) \sin(2 \pi f_o \tau) \tag{23}$$

The 'a' and 'b' factors account for the different sensitivities of the sine and cosine channels with target range. Assuming that the channel gains are adjusted so that "a" is equal to "b", the ADDER output will be:

$$E_A = a E_v \cos(2 \pi f_v t - 2 \pi f_o \tau) \tag{24}$$

It can be seen that the amplitude of E_v is independent of τ. Changes in target range show up as a phase modulation of the voice signal. As long