(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
10 December 2009 (10.12.2009)

PCT

(10) International Publication Number
WO 2009/147406 A1

(51) International Patent Classification:
G01S 13/58 (2006.01) *G01S 7/35* (2006.01)

(21) International Application Number:
PCT/GB2009/001412

(22) International Filing Date:
5 June 2009 (05.06.2009)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
0810325.1 6 June 2008 (06.06.2008) GB

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(81) Designated States *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AO, AT, AU, AZ, BA, BB, BG, BH, BR, BW, BY, BZ, CA, CH, CL, CN, CO, CR, CU, CZ, DE, DK, DM, DO, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, GT, HN, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KN, KP, KR, KZ, LA, LC, LK, LR, LS, LT, LU, LY, MA, MD, ME, MG, MK, MN, MW, MX, MY, MZ, NA, NG, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RS, RU, SC, SD, SE, SG, SK, SL, SM, ST, SV, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, ZA, ZM, ZW.

(84) Designated States *(unless otherwise indicated, for every kind of regional protection available)*: ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HR, HU, IE, IS, IT, LT, LU, LV, MC, MK, MT, NL, NO, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

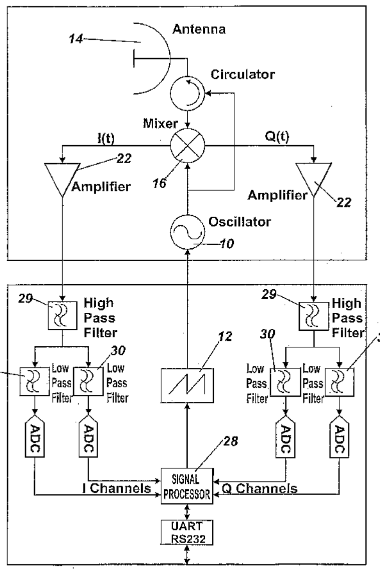
Published:

— *with international search report (Art. 21(3))*

— *before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments (Rule 48.2(h))*

(54) Title: RADAR METHODS AND APPARATUS

Fig. 3



(57) Abstract: In an FMCW radar, the signals are digitised at a low rate which is the same as, or a relatively low integral multiple of, the frequency modulation repetition rate. In this manner all the higher frequency components are folded back into a single range bin, and processing and memory requirements considerably reduced. The velocity of the moving target can be determined in the usual manner. Despite folding back to a single range bin, the range of a target is determined by splitting the return signal and feeding it into two channels having different frequency responses so that the target signal amplitude is different in the two channels. The amplitude difference is detected and used to determine the range of the target. In another embodiment, the signal is split into two channels digitised at the same digitisation rate but with a preset delay between the two channels. The relative phase difference between the relevant two target signals from the channels is determined thereby to determine the range of the target.

WO 2009/147406 A1

Radar Methods and Apparatus

This invention relates to methods and apparatus for detecting a moving target and in particular, but not exclusively, to such apparatus and methods for traffic monitoring.

For many years low cost radars have been used for traffic monitoring and speed enforcement devices. These have commonly used continuous wave (CW) radars that are only capable of measuring the speed of the target but not its range. Frequency modulated continuous wave (FMCW) radars have been used in many high bandwidth applications and offer good range resolution together with the ability to determine the range and velocity of a target. In many applications, the signal processing needs can be reduced by filtering the signal and digitising the required part of the returned signal bandwidth. However, even in these radars, the amount of signal processing required can still be significant and this generally means that a low cost radar is difficult to realise. Existing such FMCW devices still have relatively high processing and memory requirements which increase their cost and also mean that they draw significant amounts of power and so their use is usually limited to applications where there is a mains supply available.

In a typical FMCW radar employing complex sampling, the digitisation rate at which samples are measured is the number of range bins to process multiplied by the modulation repeat waveform repeat frequency MRF, otherwise referred to herein as the frequency modulation repetition rate. This ensures that all the frequencies for the targets within the wanted range for the radar are fully represented in the digital domain as required by Nyquist Shannon theorem. We have found however that by deliberately undersampling the return signal at a digitisation rate that is equal to, or a relatively low integral multiple of, the frequency modulation repetition rate, the range bins or cells are folded back onto each other, but that range data can still be determined.

We have therefore designed methods and apparatus for detecting a moving target which make use of FMCW techniques but which have considerably reduced memory and power requirements.

Accordingly, in one aspect, this invention provides a method of detecting a moving target, which comprises:

transmitting towards said target a frequency modulated radio signal at a given frequency modulation repetition rate;

receiving the return signal and digitising it at a rate equal to said frequency modulation repetition rate or a relatively low integral multiple thereof;

processing said transmitted and return signals to obtain data representative of the velocity of said target.

In specific implementations of the above method, the matching of the digitisation rate of the return signal with the repetition rate of the frequency modulation means that, in the frequency domain, the ground return signals all fold back onto the zero frequency. Also all targets within the detection range of the radar with the same velocity will fold back to the same frequency. This means that a previous requirement of several tens of fast Fourier transform (FFT) operations is now considerably reduced thereby reducing the memory and processing requirements. The term "relatively low" in relation to the integral multiple of the frequency modulation repetition rate is used to distinguish from these conventional techniques where the return signal is digitised at a rate equal to the number of range bins or cells multiplied by the frequency modulation repetition rate. Thus a relatively low rate is one which is less than double the radar's maximum range as required normally by the Nyquist-Shannon sampling theorem. A typical upper limit for the integral would be 5.

Although it would be possible to make use of such methods simply to determine the velocity of a target, in preferred embodiments the transmitted and return signals are processed to also obtain data representative of the range of said target. It will be appreciated that because the frequencies have effectively all been folded back into the first range bin, it is necessary to disambiguate the range bin signals. We have developed two preferred schemes for doing this namely an amplitude-based scheme and a phase-based scheme.

In the amplitude-based scheme, the return signal is split and fed into two channels having different frequency responses, and the data from each channel is then processed to obtain respective target signal amplitudes, with the respective target signal amplitudes then being compared to determine an estimate of the range of said target. Although various differential frequency responses are possible, in one arrangement, both channels have a high pass frequency response (similar to the range compensation filter in existing FMCW arrangements). One channel then has an additional low pass filter with a cutoff frequency of F_{low} while the other channel has an additional low pass filter with a cutoff frequency of F_{high} where $F_{low} < F_{high}$ A frequency response of a typical implementation is shown in Figure 4 of the accompanying drawings, with the solid line indicating the difference between the two responses. Thus, knowing the signal amplitudes of the folded frequency of the target detected in each channel, together with the differential frequency responses of each channel, the pre-folded frequency of the target signal can be determined, and from that the range of the target deduced.

In the phase-based scheme, the return signal is split into two channels which are digitised at the same digitisation rate but with a preset time delay between two channels, the data from the channels then being processed to determine the relative phase difference between the relevant two target signals from the channels, thereby to determine the pre-folded frequency of the target signal and thus an estimate of the range of said target.

It will be appreciated that this phase-based technique also deduces an estimate of the relevant pre-folded frequency of the target signal to determine its range. In this instance by having a slight delay between the digitisation of both channels, the phase difference can be measured and, knowing this together with the time delay, the frequency can be deduced.

The invention also extends to apparatus for detecting a moving target, which apparatus comprises:

means for transmitting towards said target of frequency modulated radio signal at a given frequency modulation repetition rate;

means for receiving the return signal;

means for digitising said return signal at a rate equal to said frequency modulation repetition rate or a relatively low integral multiple, thereof, and

means for processing said transmitted and return signals to obtain data representative of the velocity of said target.

The frequency modulation may be of any convenient form for example sawtooth or triangular waveform with a sawtooth waveform being preferred.

Whilst the invention has been described above, it extends to any inventive combination or sub-combination of the features set out above, in the following or in the following description of claims.

The invention may be performed in various ways, and two embodiments thereof will now be described by way of example only, reference being made to the accompanying drawings in which :

Figure 1 is a schematic view of a typical prior art FMCW radar;

Figure 2 is view of a typical FMCW radar mixer output in the frequency domain;

Figure 3 is a first embodiment of a radar of this invention in which range is determined by amplitude comparison;

Figure 4 is a diagram showing the filter characteristics of the filters employed in the embodiment of Figure 3;

Figure 5 is a view of a typical FMCW radar mixer output in the frequency domain from the embodiment of Figure 3;

Figure 6 is a second embodiment of FMCW radar in accordance with this invention in which range is determined by phase comparison, and

Figure 7 is a flow chart showing operation of an example of the phase comparison embodiment of Figure 6.

Referring initially to Figure 1, in a typical known FMCW radar, an oscillator 10 is frequency modulated by a sawtooth waveform 12 to provide a "chirp" signal which is coupled to an antenna 14 for transmission. The received signal is supplied to a mixer 16 to be mixed with the chirp signal to provide an intermediate signal whose real and imaginary parts I(t) and Q(t) are processed through inphase and quadrature paths 18, 20 respectively. In each path the signal is amplified by an amplifier 22 to then pass through a high pass filter 24 which acts a range compensation filter, with the filtered output then passing to an analogue to digital converter 26 and the digitised signals then being supplied to a signal processor 28. The range compensation filter is designed so that signals from all targets have a similar level. Objects further away give rise to higher frequencies but their smaller size due to range means that the amplitude of the return signal decreases with distance and so the high pass filter compensate for this.

Figure 2 shows a typical frequency spectrum of a typical signal from the mixer 16. Each peak in the frequency spectrum represents a potential target. From the spectrum it can be seen that there is a signal that repeats every 25kHz. These are signals due to reflections from the ground (zero velocity). In many instances these signals are not of interest as in many radar applications it is only moving targets that are of interest. Thus it is the signal peaks between these ground returns that are the wanted moving target signals. To process this signal the analogue to digital converters 26 would typically be required to digitise at a rate of 400 kilo samples per second (KSPS). The signal processor 28 typically applies a series of range FFTs followed by a series of Doppler FFTs to extract the target information. The following FFTs would have to be carried out in this example which would have 64 Doppler bins:

- 64* 16 point range FFTs

- 16* 64 point Doppler FFTs

This amounts to a huge number of data samples which means that a large amount of memory and significant processing power is required.

Turning now to the embodiments described below, in each of these the signals are digitised at a much lower rate than is usual in an FMCW radar with range measuring capability. In the examples given, the signals are digitised at 25KSPS instead of 400 KSPS. These signals are therefore being decimated by a factor of 16. This means that the analogue signal is heavily undersampled and when the digitised signals are FFT'd only a 25KHz bandwidth of signal can be seen. All the higher frequency components from the original bandwidth of 400KHz will have been folded back into the 25KHz bandwidth. The digitisation frequency has been chosen so that all the ground return signals fold back to 0Hz. This is achieved by choosing the frequency modulation repetition rate to be the same as the digitisation rate. In these embodiments, which we refer to as decimating FMCW radars, the moving target can be clearly seen in the frequency domain but its range can no longer be determined by the conventional technique. We describe two different embodiments in which the range can be determined. The first embodiment relies on an amplitude comparison step, and the second embodiment relies on a phase comparison step. Both embodiments essentially determine the original frequency of the target signal in the pre-folded spectrum. We refer to "pre-folded" and "folded" frequencies to differentiate between the signals before and after the processing which causes the folding.

Turning firstly to the amplitude comparison decimating FMCW radar shown in Figure 3, components similar to those used in the conventional arrangement of Figure 1 are given like reference numerals. As previously, a frequency modulation derived at 12 is applied to an oscillator 10 to provide a transmitted signal which is passed to the antenna 14. The return signal is mixed with the transmitted signal to obtain an intermediate signal which is processed along inphase and quadrature paths 18, 20. However, in each path, the signals are split, passed to two high pass filters 29 and then fed into two sets of low pass filters 30,32 having different cut off frequencies F_{Low} and F_{High} respectively. It should be noted that it is not necessary to have a high pass filter response, indeed in some instances it may be preferred not to provide the range compensation function so as to make closer targets, represented by low frequencies, have a much larger return signal. The low pass filter with the lower cutoff frequency (F_{Low}) in the example given discriminates in favour of targets that are closer. The filter responses should be selected so that they allow the pre-folded frequency of the target to be determined.

The output of the mixer in the frequency domain is shown in Figure 4 from which it will be seen that the presence of a target is evident from the signal peak at 36. The frequency at which this peak appears in Figure 4 indicates the velocity of the target. In order to determine the range of the target, the filtered channels are FFTed and similar spectrums are observed. However, because the low pass filters have different cutoff frequencies their output target signal amplitudes are different and the extent of the difference is indicative of the pre-folded frequency of the target return and thus is used to determine its range.

In the above arrangement, the number of FFTs required has been reduced to just two 64 point FFTs followed by an amplitude comparison step to determine the targets range. This arrangement therefore considerably reduces the amount and speed of processing required and therefore allows a relatively inexpensive processor to be used.

Referring now to the phase comparison embodiment, in this arrangement the FMCW radar is modified as shown in Figure 6. Similar components are given similar references. For this type of radar there is no additional filtering required for the extra channel. Instead, in this arrangement, the signal received from the high pass filter 24 is digitised at a low rate but, split into channels that are clocked by a clock generator 34 at the same frequency, but with one slightly delayed. When the two channels are digitised, the amplitude response will be similar. However, because of the slight time difference introduced between digitisation of the channels there will be a phase difference between the slightly delayed samples of the target return. The phase difference measured is dependant on the pre-folded frequency of the target return that caused the FFT output. The higher the original pre-folded frequency, the greater the phase difference measured will be, i.e.:

$$\text{Channel1}(t) = \sin(\omega t)$$

$$\text{Channel2}(t) = \sin(\omega(t+\Delta t))$$

$$\text{Phase difference} = \omega\Delta t$$

Therefore knowing the phase difference and the time delay, the pre-folded frequency (ω) can be determined which in turn identifies the range of the target. This phase difference method has the advantage that it is not necessary to calibrate or match the filters used in the amplitude-based method. Two options are available; firstly the delay between the samples may be small in terms of the period of the pre-folded frequency of the target return i.e. equivalent to a phase difference of a few degrees, or it may be much larger, approaching half the period of the sample frequency. A benefit of this latter arrangement is that it means that the analogue to digital converter has longer to carry out digitisation because the samples are more evenly distributed timewise.

A more detailed mathematical analysis will now be made of the FMCW radar in accordance with preferred embodiments of this invention, and the phase comparison method. In a FMCW radar the received signal from a target is downconverted from its carrier frequency using the transmitted waveform. Using this technique the lower frequencies in the downconverted signal represent close targets and higher frequencies represent targets at larger ranges. If the user only wishes to observe close targets it is therefore possible to filter the lower frequencies so that only a small bandwidth needs to be processed. Because of this the user can transmit a very large bandwidth of say 100's of MHz but only needs to process say 10MHz for the ranges the user wishes to process.

In the technique according to the invention, the transmitted waveform is frequency modulated using a sawtooth waveform with a modulation repeat frequency of MRF. The downconverted baseband signal (Rx(t)) for a target return signal can be represented in complex form by:

$$Rx(t) = e^{2\pi i \cdot \left(R \cdot MRF + F_D\right) \cdot t}$$

in which:

R is the range bin of a target and is an integral number, MRF is the modulation waveform repeat frequency, and

F_D is the Doppler shift of the signal due to the movement of the target.

This signal is digitised in a radar by a analogue to digital converter, ADC, that samples the waveform at a sample frequency of F_S. The time at which each sample is measured is :

$$t = \frac{n}{F_S},$$

where n is the sample number and is an integral number.

The value of the nth sample of the waveform is given by

$$R(n) = e^{2\pi i \cdot (R \cdot MRF + F_D) \cdot \frac{n}{F_S}}.$$

Normally the rate at which samples are measured is the number of range bins to process, multiplied by the MRF. This ensures that all the frequencies for the targets within the wanted range for the radar at fully represented in the digital domain as required by Nyquist Shannon theorem.

In the present technique, by setting F_S to MRF, the data is being undersampled but the sample rate F_S has been significantly reduced. The digitised waveform for this case is:

$$Rx(n) = e^{2\pi i \cdot (R \cdot MRF + F_D) \cdot \frac{n}{MRF}}$$

which can be written $Rx(n) = e^{\frac{2\pi i \cdot R \cdot MRF \cdot n}{MRF} + \frac{2 \cdot \pi \cdot i F_D \cdot n}{MRF}}$

Because n and R are integers and $e^{2im\pi} = 1$, (where m is an integral number) this becomes

$$Rx(n) = e^{2\cdot\pi\cdot i\cdot R\cdot n} \cdot e^{\frac{2\cdot\pi\cdot i F_D\cdot n}{MRF}}$$ which simplifies to $$Rx(n) = e^{\frac{2\cdot\pi\cdot i F_D\cdot n}{MRF}}$$

From this it can be seen that by digitising at the modulation repeat frequency, MRF, the digitised received signal will always have a frequency component that is the Doppler frequency component of the target no matter the range of the target. Therefore by performing a fast Fourier transform on this data set the Doppler component of a target can be determined and therefore its velocity has been measured.

Turning now to the phase comparison method, in this method of range finding the phase difference information is used between two sampled data sets that are sampled Δn apart where Δn is a fraction of n.

The first sample set for a downconverted target signal is represented by

$$Rx1(n) = e^{2\pi i\cdot\left(R\cdot MRF + F_D\right)\cdot\frac{n}{F_S}}$$

The second sample set sampled Δn from the first sample set, Rx1(n), is given by

$$Rx2(n) = e^{2\pi i\cdot\left(R\cdot MRF + F_D\right)\cdot\frac{n+\Delta n}{F_S}}$$

Again if the sample rate is F_S = MRF, the first sample set is now given by

$$Rx1(n) = e^{\frac{2\cdot\pi\cdot i F_D\cdot n}{MRF}}$$ and the second sample set is given by

$$Rx2(n) = e^{2\pi\cdot i\cdot\left(R\cdot MRF + F_D\right)\cdot\frac{n+\Delta n}{MRF}};$$ $$Rx2(n) = e^{\frac{2\pi\cdot i\cdot R\cdot MRF(n+\Delta n)}{MRF} + \frac{2\cdot\pi\cdot i F_D(n+\Delta n)}{MRF}};$$

$$Rx2(n) = e^{2\pi \cdot i \cdot R \cdot (n+\Delta n) + \frac{2\cdot\pi\cdot i F_D (n+\Delta n)}{MRF}} ; \qquad Rx2(n) = e^{2\pi \cdot i \cdot R \cdot n + 2\cdot\pi\cdot i \cdot R\Delta n + \frac{2\cdot\pi\cdot i F_D (n+\Delta n)}{MRF}}$$

But R and n are integers, and so $e^{2\pi \cdot i \cdot R \cdot n} = 1$

Therefore Rx2(n) can be rewritten as

$$Rx2(n) = e^{2\cdot\pi\cdot i \cdot R\Delta n + \frac{2\cdot\pi\cdot i F_D (n+\Delta n)}{MRF}} ; \qquad Rx2(n) = e^{2\cdot\pi\cdot i \cdot R\Delta n + \frac{2\cdot\pi\cdot i F_D \cdot n}{MRF} + \frac{2\cdot\pi\cdot i F_D \cdot \Delta n}{MRF}}$$

The sample sets Rx1(n) and Rx2(n) can be processed to determine their frequency domain characteristics using various processing techniques of which the fast Fourier transform is one. Using a fast Fourier transform the amplitude and phase of the various frequency components can be determined. By using the phase information in Rx1(n) and Rx2(n) a method will now be developed to determine the range of a target.

By dividing Rx2(n) by Rx1(n) the resultant phase component gives the phase difference between the data sets.

$$\frac{Rx2(n)}{Rx1(n)} = \frac{e^{2\cdot\pi\cdot i \cdot R\Delta n + \frac{2\cdot\pi\cdot i F_D \cdot n}{MRF} + \frac{2\cdot\pi\cdot i F_D \cdot \Delta n}{MRF}}}{e^{\frac{2\cdot\pi\cdot i F_D \cdot n}{MRF}}} ; \qquad \frac{Rx2(n)}{Rx1(n)} = e^{2\cdot\pi\cdot i \cdot R \cdot \Delta n + \frac{2\cdot\pi\cdot i F_D \cdot \Delta n}{MRF}}$$

$$\frac{Rx2(n)}{Rx1(n)} = e^{2\cdot\pi\cdot i\cdot\Delta n\left(R+\frac{F_D}{MRF}\right)}$$

The phase difference between the data sets is therefore

$$PhDif = 2\cdot\pi\cdot\Delta n\left(R + \frac{F_D}{MRF}\right)$$

From this it can be seen that the phase difference between the two data sets is dependant on the Doppler frequency of the target, F_D and the range bin the target is in, R. The Doppler frequency of the target can be determined from either data set, using for instance a fast Fourier transform, and therefore the range R can be determined. However, there are restrictions because phase can only be measured from 0 to 360 degrees. The accuracy with which the phase difference can be measured determines the number of ranges that can be measured or maximum number of range cells, and so:

Maximum Number of Range Cells = 360 / Phase Accuracy

Now Δn needs to be chosen to give useful results. If M is the number of range cells the radar is to measure, then Δn is chosen as.

$$\Delta n = \frac{1}{M}$$

Therefore the phase difference between the two data sets is given by the phase of the equation below.

$$\frac{Rx2(n)}{Rx1(n)} = e^{\frac{2\cdot\pi\cdot i}{M}\cdot\left(R+\frac{F_D}{MRF}\right)}$$

Thus the phase difference is

$$\mathrm{PhDif} = \frac{2\cdot\pi}{M}\cdot\left(R + \frac{F_D}{MRF}\right) \qquad \mathrm{PhDif} = \frac{2\cdot\pi\cdot R}{M} + \frac{2\cdot\pi\cdot F_D}{M\cdot MRF}$$

The maximum Doppler frequency is usually chosen to be half the MRF in a radar design.

Therefore $\frac{F_D}{MRF} < 0.5$

In this case the error caused by the Doppler of the target will only cause a maximum error of half a range cell. This is acceptable and therefore the Doppler frequency component may be ignored, which gives:

$$\mathrm{PhDif} = \frac{2\cdot\pi\cdot R}{M}$$

When $\Delta n = 1/M$, the time between data samples measured for each data set is small. A analogue to digital converter, ADC, usually has a minimum sample time and therefore may not be able to sample data this close in time. To avoid using a high specification ADC that can sample this closely a second ADC can be used instead to capture the second sample set. Both ADCs would then be clocked at the MRF rate.

Figure 7 shows the processing steps for the above technique.

An alternative would be to use a single low specification ADC converter if the samples required could be sampled at times so as to make the interval between samples approximately equal.

Therefore choose Δn to be : $\Delta n = \frac{1}{2} - \frac{1}{M}$

Again Rx2(n)is divided by Rx1(n) to determine the phase difference

$$\mathrm{PhDif} = 2\cdot\pi\cdot\Delta n\left(R + \frac{F_D}{MRF}\right); \qquad \mathrm{PhDif} = 2\cdot\pi\cdot\left(0.5 - \frac{1}{M}\right)\cdot\left(R + \frac{F_D}{MRF}\right);$$

$$\mathrm{PhDif} = \left(\pi - \frac{2\pi}{M}\right)\cdot\left(R + \frac{F_D}{MRF}\right); \qquad \mathrm{PhDif} = \left(\pi - \frac{2\pi}{M}\right)\cdot R + \left(\pi - \frac{2\pi}{M}\right)\cdot\frac{F_D}{MRF}.$$

In this case the term $\left(\pi - \frac{2\pi}{M}\right)\cdot\frac{F_D}{MRF}$ is significant as it can be > $2\pi/M$

So in this case the Doppler component would have to be compensated for. To compensate for the Doppler frequency the data set is fast Fourier transformed to determine the Doppler frequency and this is then substituted back into the above equation so that the range, R, can then be determined.

CLAIMS

1. A method of detecting a moving target, which comprises:

transmitting towards said target a frequency modulated radio signal at a given frequency modulation repetition rate;

receiving the return signal and digitising it at a rate equal to said frequency modulation repetition rate or a relatively low integral multiple thereof;

processing said transmitted and return signals to obtain data representative of the velocity of said target.

2. A method according to Claim 1, wherein the transmitted and return signals are processed to also obtain data representative of the range of said target.

3. A method according to Claim 1 or Claim 2, wherein the return signal is split and fed into two channels having different frequency amplitude responses, and the data from each channel is then processed to obtain respective target signal amplitudes, with the respective target signal amplitudes then being used to determine an estimate of the range of said target.

4. A method according to claim 3 wherein analogue filtering is implemented in such a way so that closer targets, with the same radar cross section, have a larger signal amplitude and therefore biasing the radar towards detecting the nearest target to the radar.

5. A method according to Claim 1 or Claim 2, wherein the return signal is digitised into two separate data sets which are digitised at the same digitisation rate but with a preset time delay between the two data sets, the data from the channels then being processed to determine the relative phase

difference between the relevant target signals in the data sets, thereby to determine the pre-folded frequency of the target signal and thus an estimate of the range of said target.

6. Apparatus for detecting a moving target, which apparatus comprises:

means for transmitting towards said target of frequency modulated radio signal at a given frequency modulation repetition rate;

means for receiving the return signal;

means for digitising said return signal at a rate equal to said frequency modulation repetition rate or a relatively low integral multiple thereof, and

means for processing said transmitted and return signals to obtain data representative of the velocity of said target.

7. A method substantially as hereinbefore described with reference to any of Figures 3 to 5.

8. Apparatus substantially as hereinbefore described and illustrated with reference to any of Figures 3 to 5.

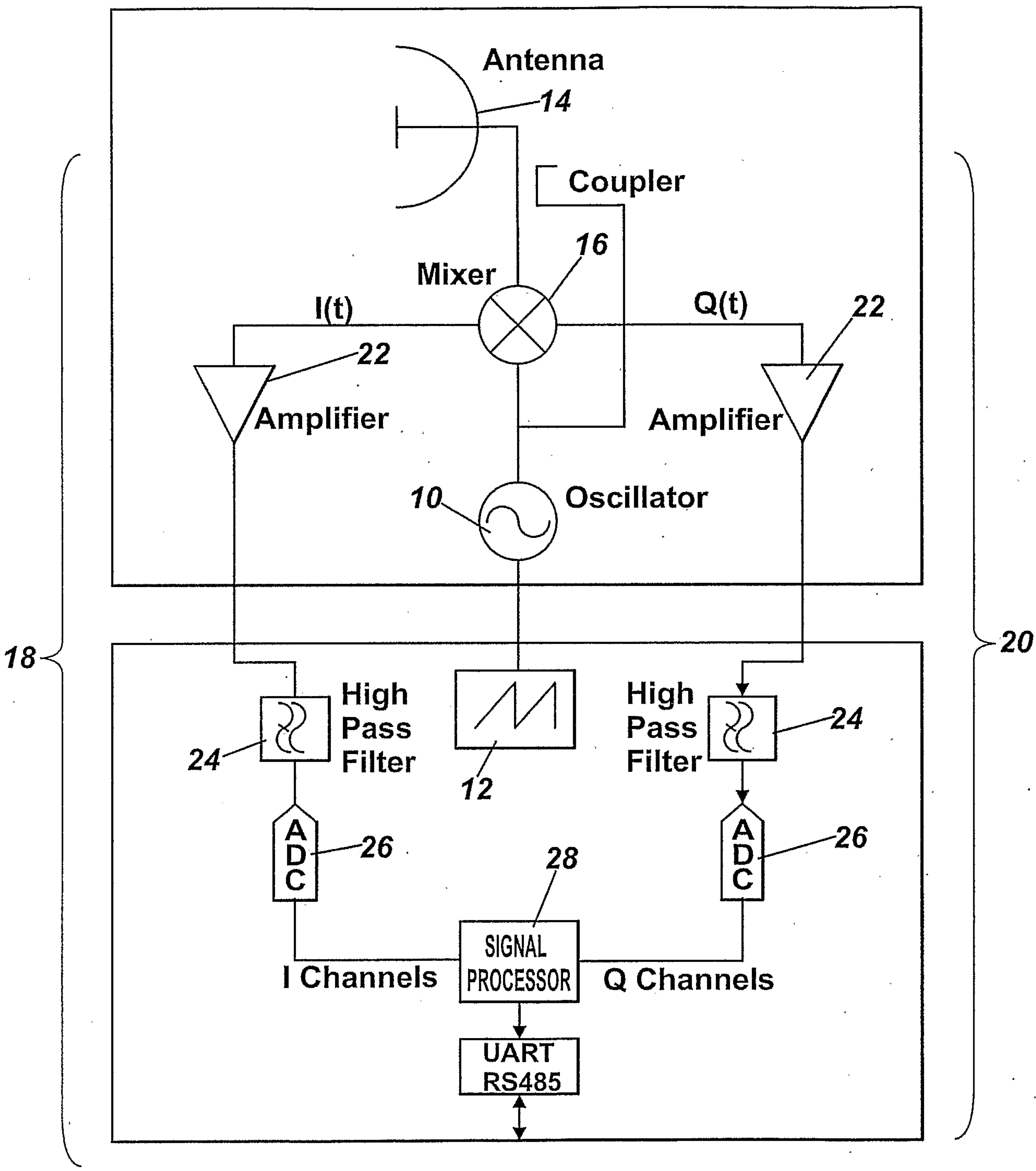

Antenna
14
Coupler
16
Mixer
I(t)
Q(t)
22
22
Amplifier
Amplifier
10
Oscillator
18
20
High
Pass
Filter
24
12
High
Pass
Filter
24
ADC
26
28
ADC
26
SIGNAL
PROCESSOR
I Channels
Q Channels
UART
RS485

Fig. 1

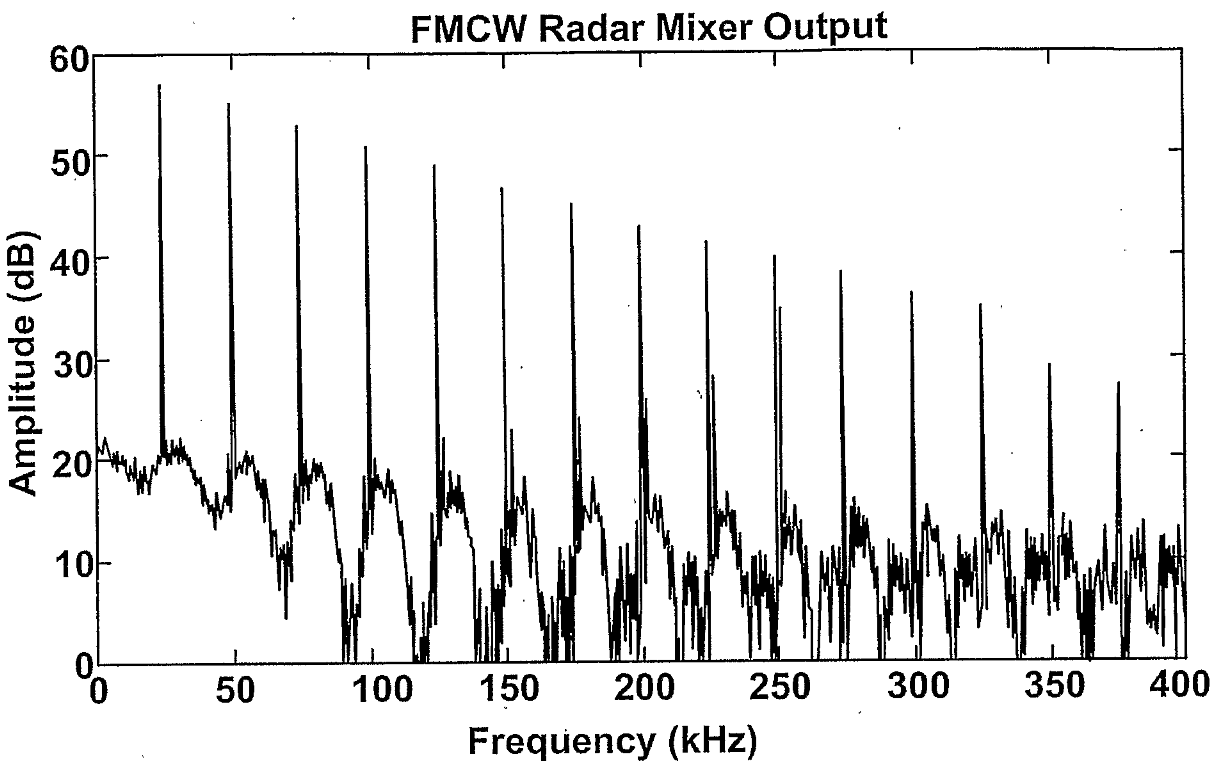

FMCW Radar Mixer Output
Amplitude (dB)
0
10
20
30
40
50
60
0
50
100
150
200
250
300
350
400
Frequency (kHz)

Fig. 2

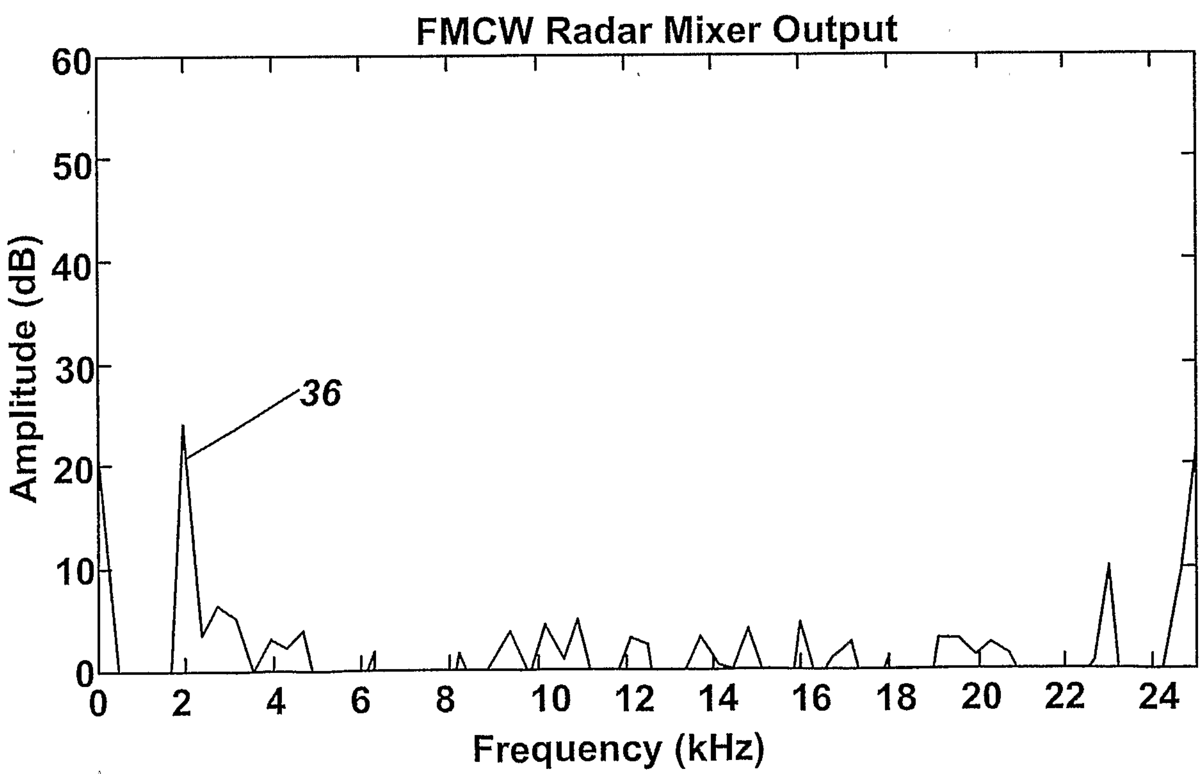

FMCW Radar Mixer Output
Amplitude (dB)
0
10
20
30
40
50
60
36
0
2
4
6
8
10
12
14
16
18
20
22
24
Frequency (kHz)

Fig. 5

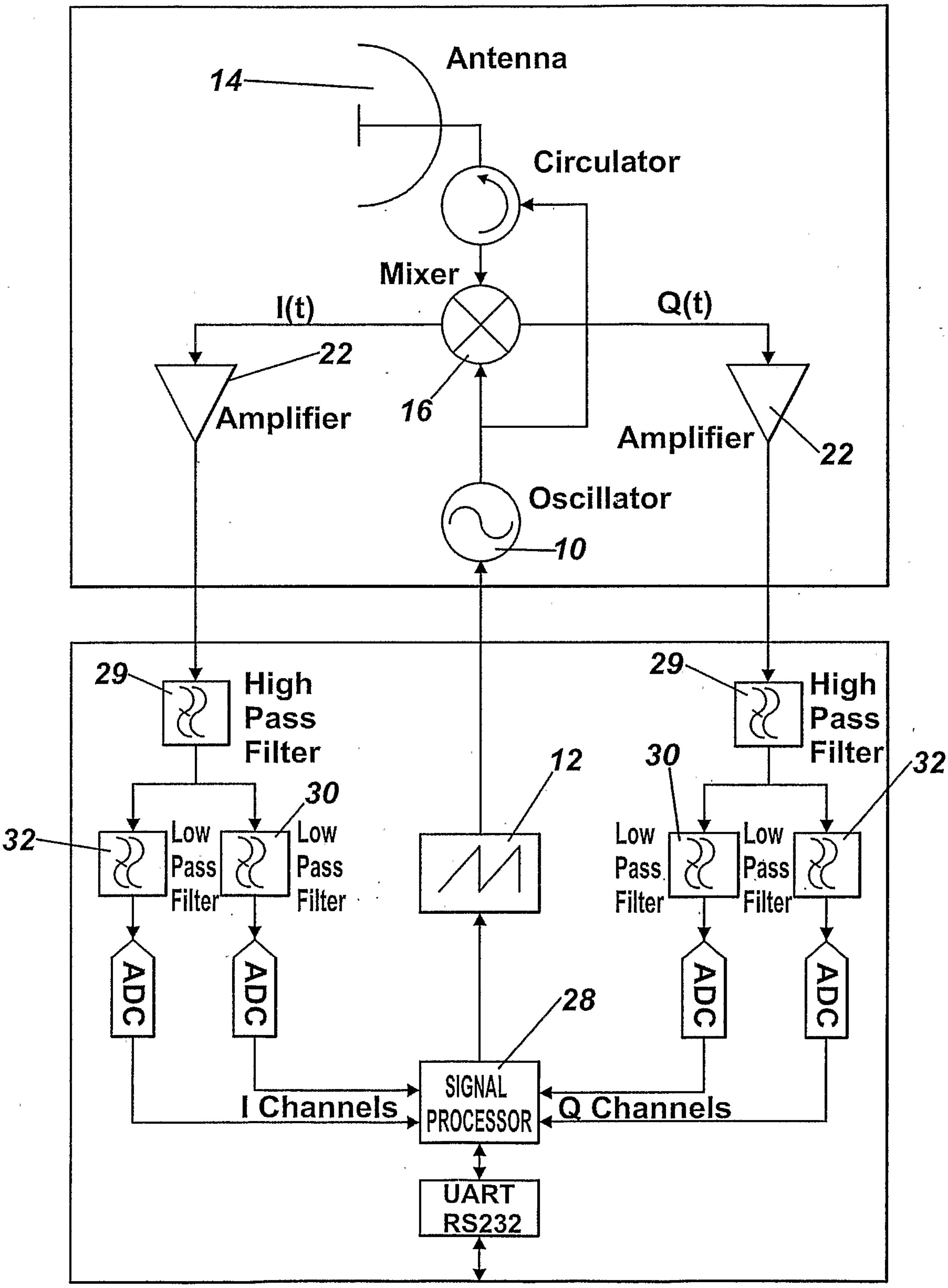

Antenna
14
Circulator
Mixer
I(t)
Q(t)
22
Amplifier
16
Amplifier
22
Oscillator
10
29
High Pass Filter
29
High Pass Filter
12
30
32
30
32
Low Pass Filter
Low Pass Filter
Low Pass Filter
Low Pass Filter
ADC
ADC
ADC
ADC
28
SIGNAL PROCESSOR
I Channels
Q Channels
UART RS232

Fig. 3

4/6

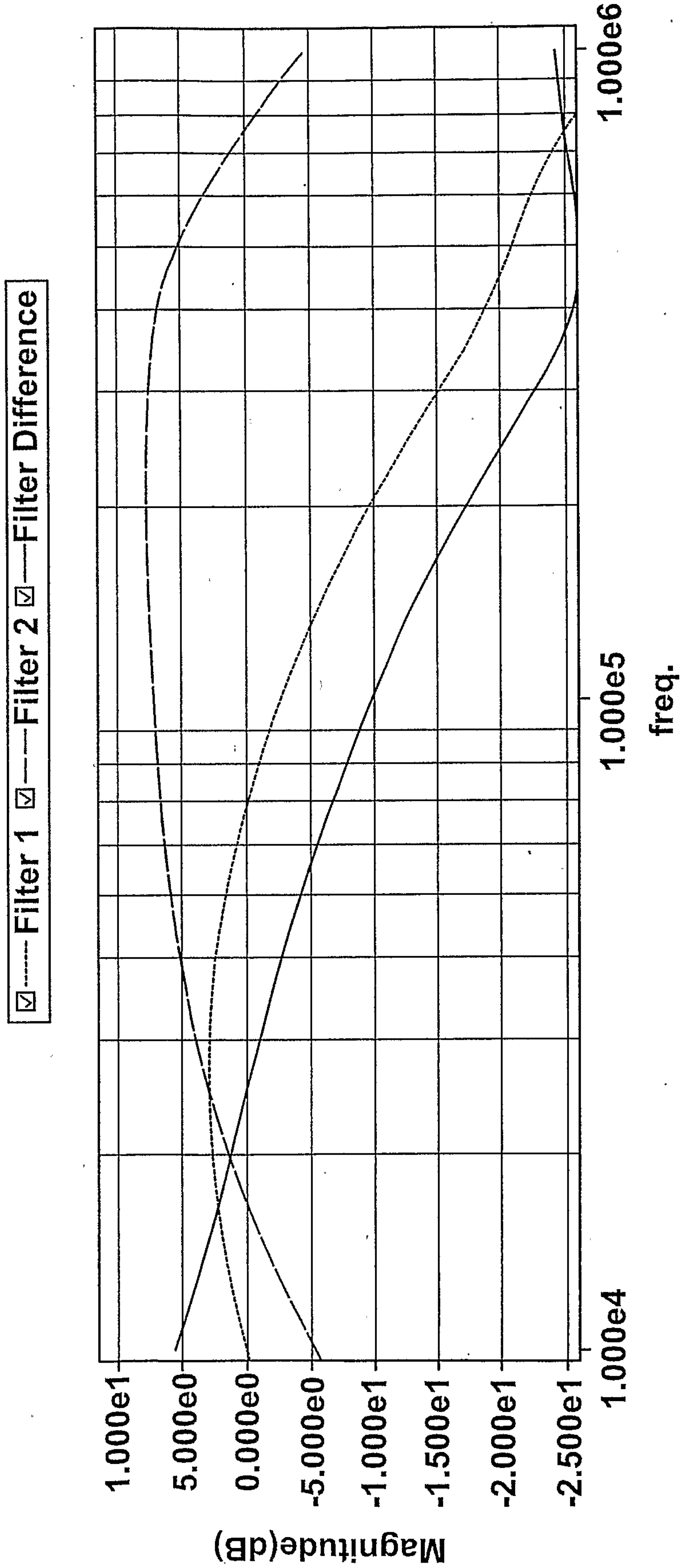

Filter 1
Filter 2
Filter Difference
1.000e1
5.000e0
0.000e0
-5.000e0
-1.000e1
-1.500e1
-2.000e1
-2.500e1
Magnitude(dB)
1.000e4
1.000e5
1.000e6
freq.

Fig. 4

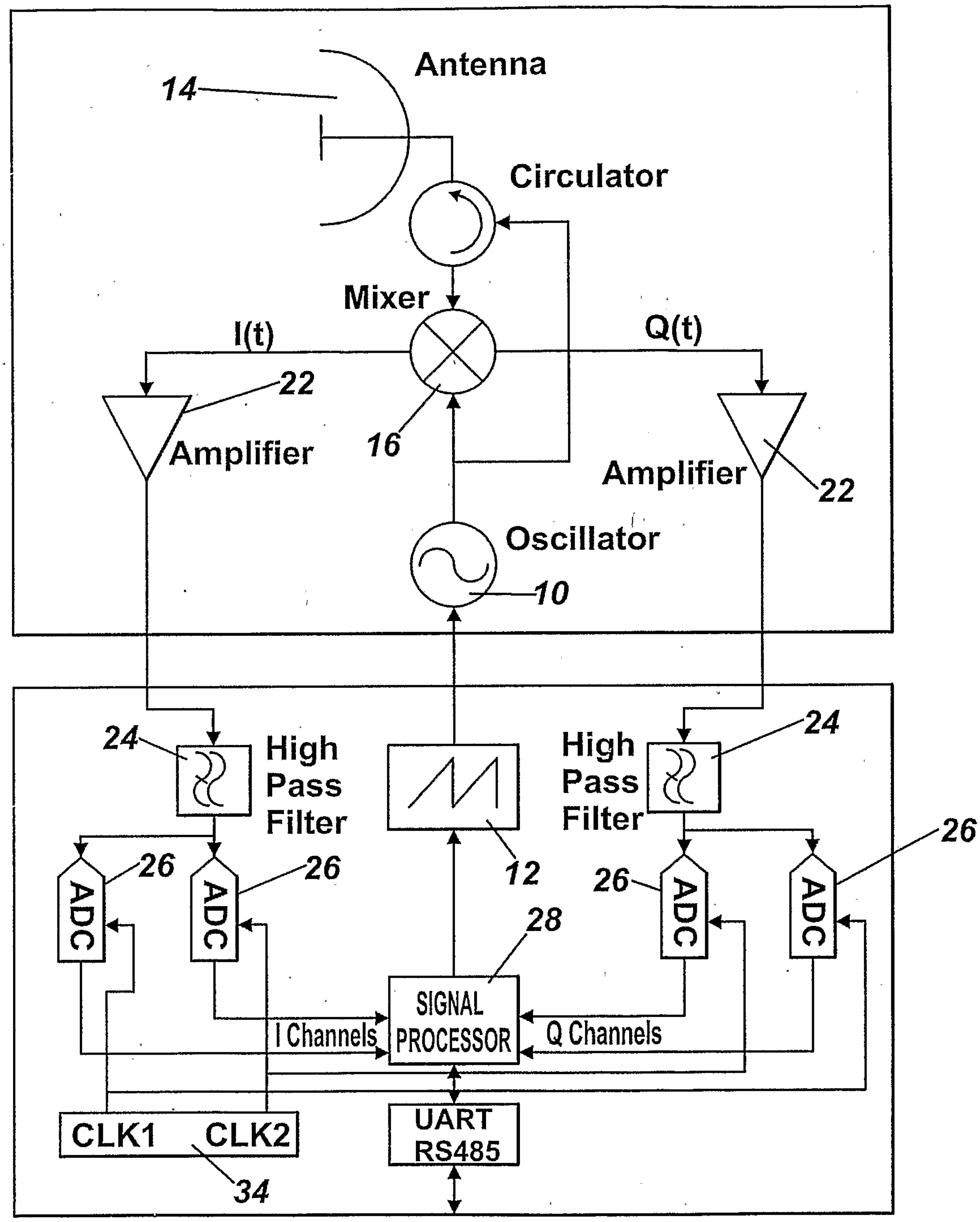

Antenna
14
Circulator
Mixer
I(t)
Q(t)
22
Amplifier
16
Amplifier
22
Oscillator
10
24
High
Pass
Filter
High
Pass
Filter
24
26
26
12
26
26
ADC
ADC
ADC
ADC
28
SIGNAL
PROCESSOR
I Channels
Q Channels
CLK1
CLK2
UART
RS485
34

Fig. 6

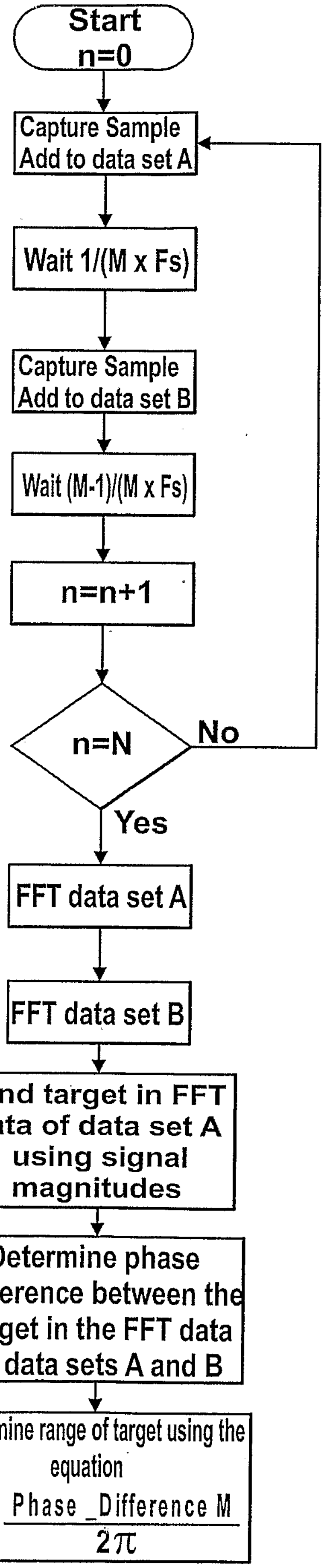

Start
n=0
Capture Sample
Add to data set A
Wait 1/(M x Fs)
Capture Sample
Add to data set B
Wait (M-1)/(M x Fs)
n=n+1
n=N
No
Yes
FFT data set A
FFT data set B
Find target in FFT data of data set A using signal magnitudes
Determine phase difference between the target in the FFT data of data sets A and B
Determine range of target using the equation
R= Phase _Difference M / 2π

Fig. 7

INTERNATIONAL SEARCH REPORT

International application No
PCT/GB2009/001412

A. CLASSIFICATION OF SUBJECT MATTER
INV. G01S13/58 G01S7/35

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
G01S

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 6 229 474 B1 (UEHARA NAOHISA [JP]) 8 May 2001 (2001-05-08) figure 5 column 1, line 5 - line 13 column 5, line 60 - line 65 -----	1-8
Y	WO 2005/093462 A (IDS INGEGNERIA DEI SISTEMI S P [IT]; MANACORDA GUIDO [IT]; MINIATI MAR) 6 October 2005 (2005-10-06) figure 1 page 2, line 15 - line 35 -----	1-8
A	US 2003/120443 A1 (SAITOU TAKUYA [JP] ET AL) 26 June 2003 (2003-06-26) figure 6 abstract ----- -/--	1,6

[X] Further documents are listed in the continuation of Box C.

[X] See patent family annex.

* Special categories of cited documents :

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"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
25 September 2009	02/10/2009

Name and mailing address of the ISA/	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Fax: (+31-70) 340-3016	Alberga, Vito

1

INTERNATIONAL SEARCH REPORT

International application No
PCT/GB2009/001412

C(Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 6 121 915 A (COOPER GEORGE R [US] ET AL) 19 September 2000 (2000-09-19) figure 1 column 2, line 5 - line 12 -----	1,6

1

INTERNATIONAL SEARCH REPORT

International application No
PCT/GB2009/001412

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
US 6229474	B1	08-05-2001	JP 3623128	B2	23-02-2005
			JP 2000338229	A	08-12-2000
WO 2005093462	A	06-10-2005	NONE		
US 2003120443	A1	26-06-2003	JP 2003185683	A	03-07-2003
US 6121915	A	19-09-2000	NONE		