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(54) **MULTI-TAP, DIGITAL-PULSE-DRIVEN MIXER**

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(52) **U.S. Cl.** ..... **375/316**; 375/345; 455/334

(58) **Field of Search** ..... 375/316, 345; 455/234, 334; 341/143

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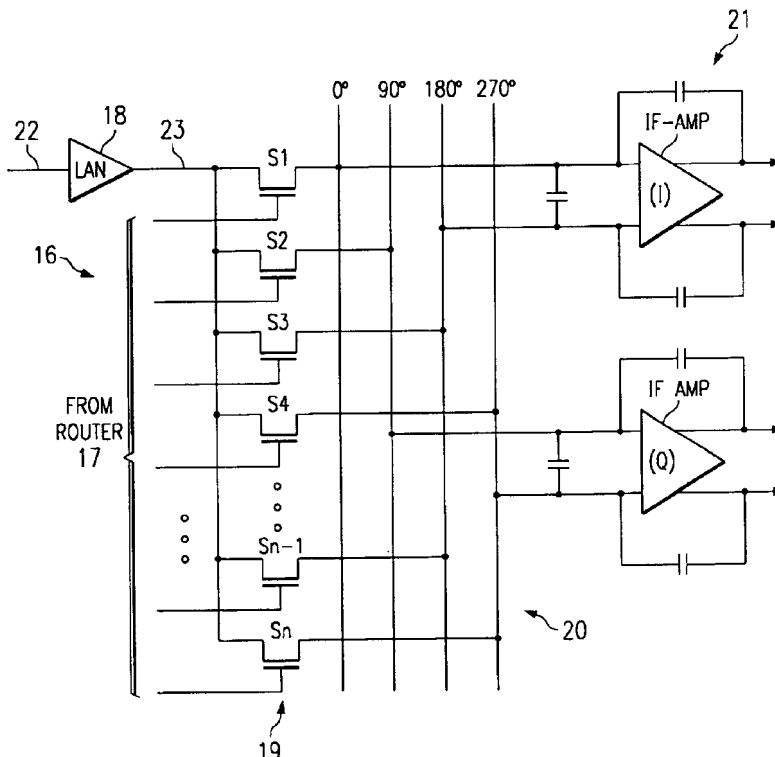
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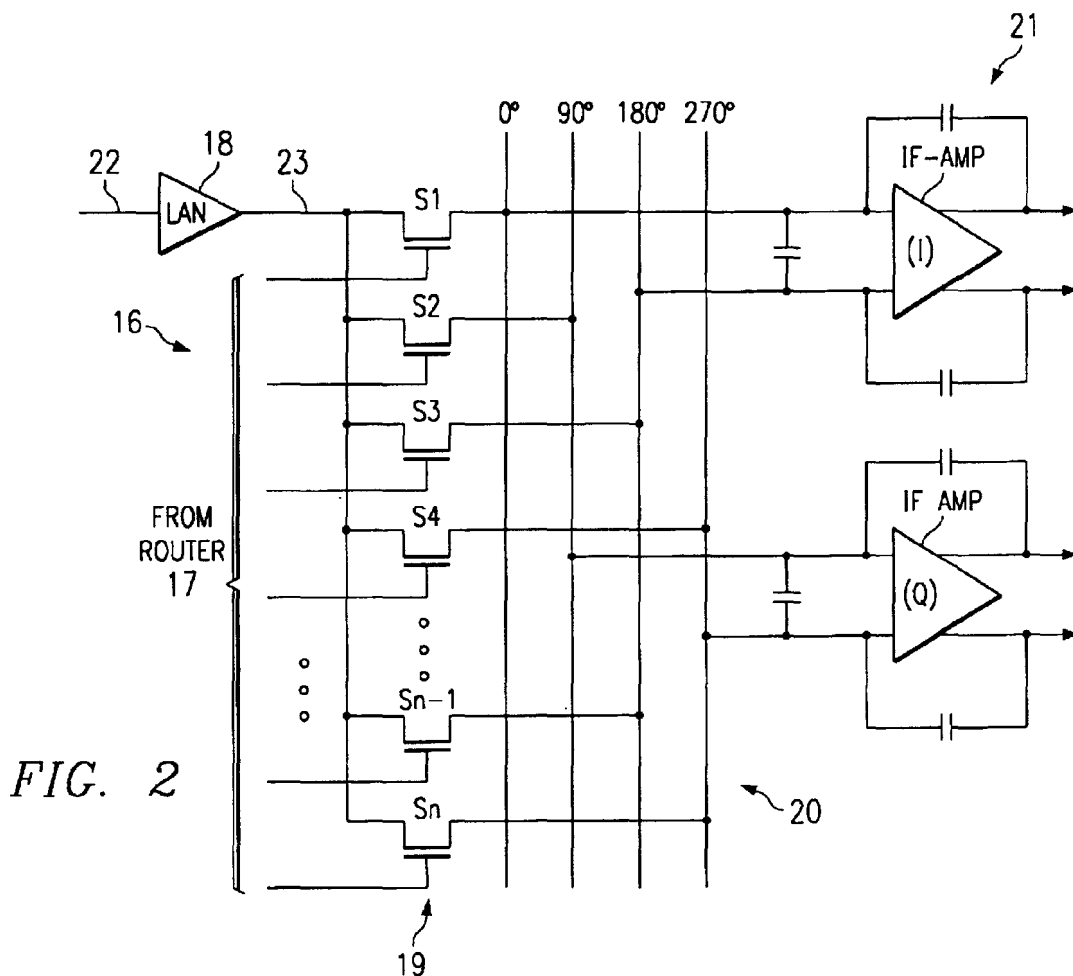
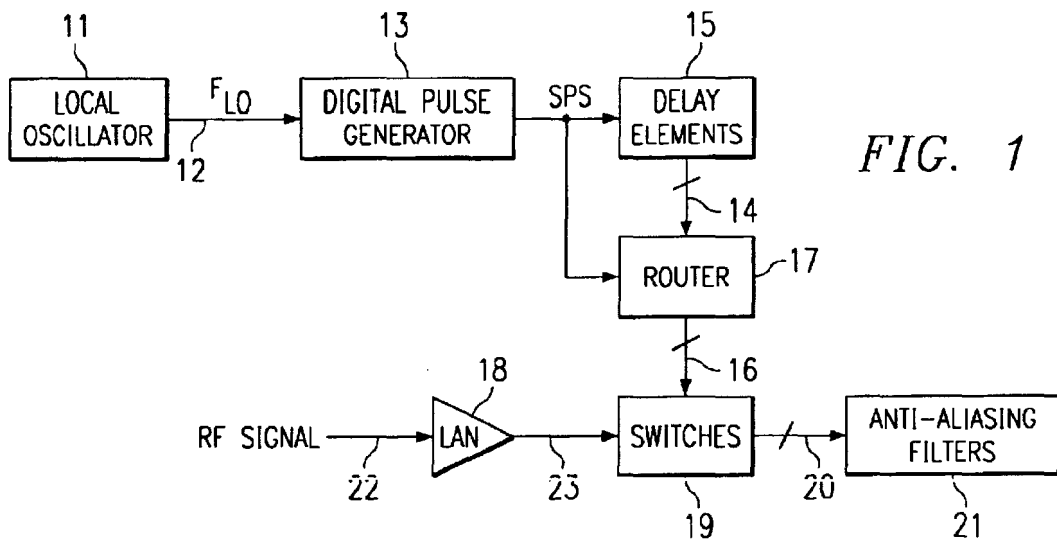
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(57) **ABSTRACT**

A multi-tap, digital-pulse-driven mixer advantageously avoids local oscillator (11) leakage by shifting the local oscillator frequency ( $F_{LO}$ ) out of the received frequency band. Low noise figures are advantageously realized by the use of digital pulses (51, 52) as mixer drive signals (16).

**28 Claims, 5 Drawing Sheets**





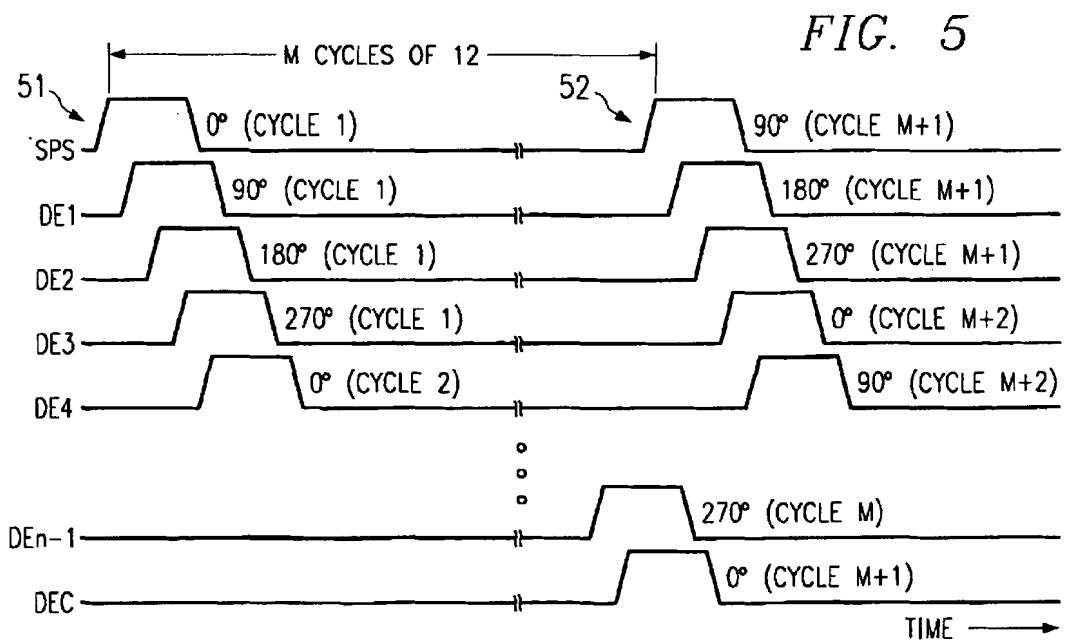
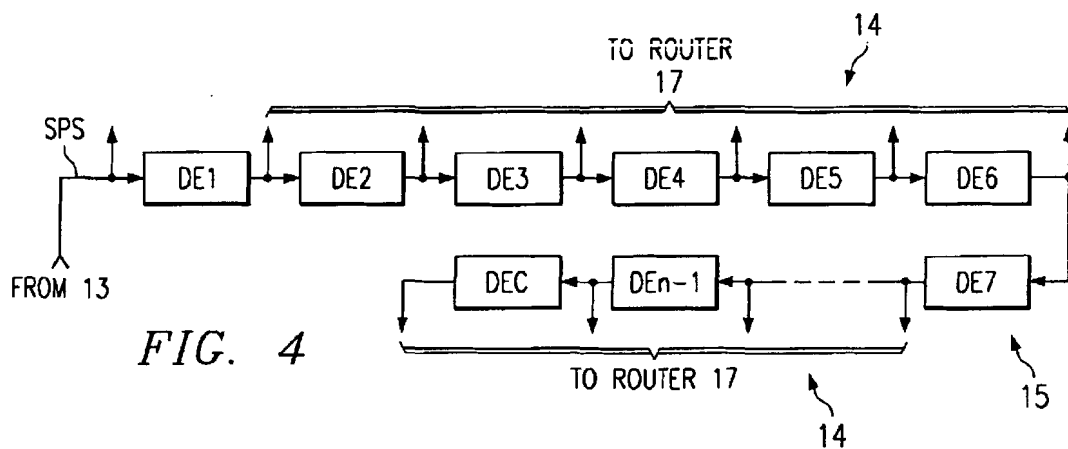
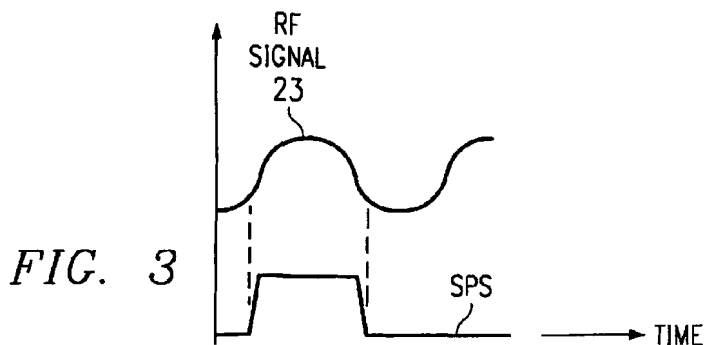
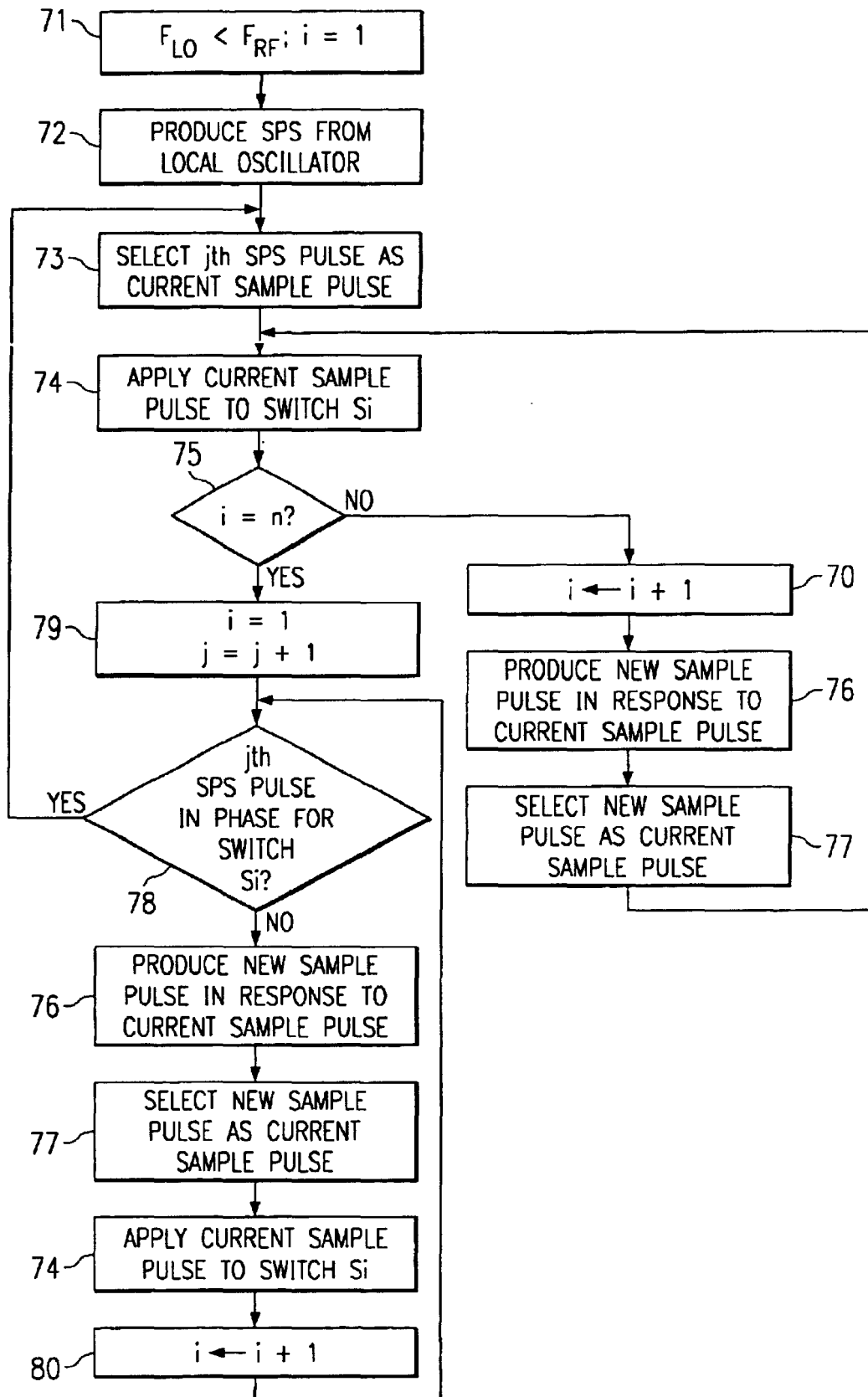


FIG. 6

SWITCH (PHASE)	RF CYCLE	PULSE SOURCE	RF CYCLE	PULSE SOURCE	RF CYCLE	PULSE SOURCE	RF CYCLE	PULSE SOURCE
S1 (0)	K	SPS	K+4	DEC	K+8	DE15	K+12	DE14
S2 (90)	↓	DE1	↓	SPS	↓	DEC	↓	DE15
S3 (180)	↓	DE2	↓	DE1	↓	SPS	↓	DEC
S4 (270)	↓	DE3	↓	DE2	↓	DE1	↓	SPS
S5 (0)	K+1	DE4	K+5	DE3	K+9	DE2	K+13	DE1
S6 (90)	↓	DE5	↓	DE4	↓	DE3	↓	DE2
S7 (180)	↓	DE6	↓	DE5	↓	DE4	↓	DE3
S8 (270)	↓	DE7	↓	DE6	↓	DE5	↓	DE4
S9 (0)	K+2	DE8	K+6	DE7	K+10	DE6	K+14	DE5
S10 (90)	↓	DE9	↓	DE8	↓	DE7	↓	DE6
S11 (180)	↓	DE10	↓	DE9	↓	DE8	↓	DE7
S12 (270)	↓	DE11	↓	DE10	↓	DE9	↓	DE8
S13 (0)	K+3	DE12	K+7	DE11	K+11	DE10	K+15	DE9
S14 (90)	↓	DE13	↓	DE12	↓	DE11	↓	DE10
S15 (180)	↓	DE14	↓	DE13	↓	DE12	↓	DE11
S16 (270)	↓	DE15	↓	DE14	↓	DE13	↓	DE12
	L → TO K+4		L → TO K+8		L → TO K+12			

TO CYCLE K ← K=K+16 ←

FIG. 7



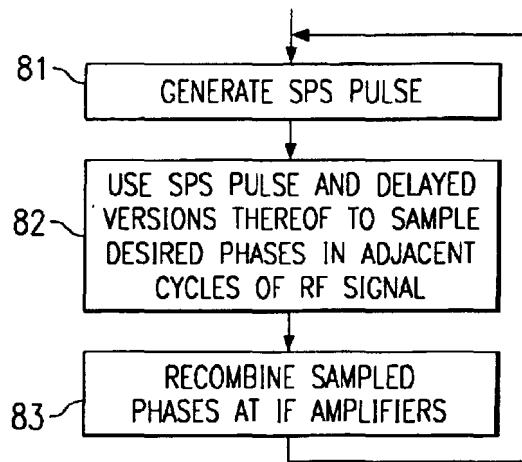


FIG. 8

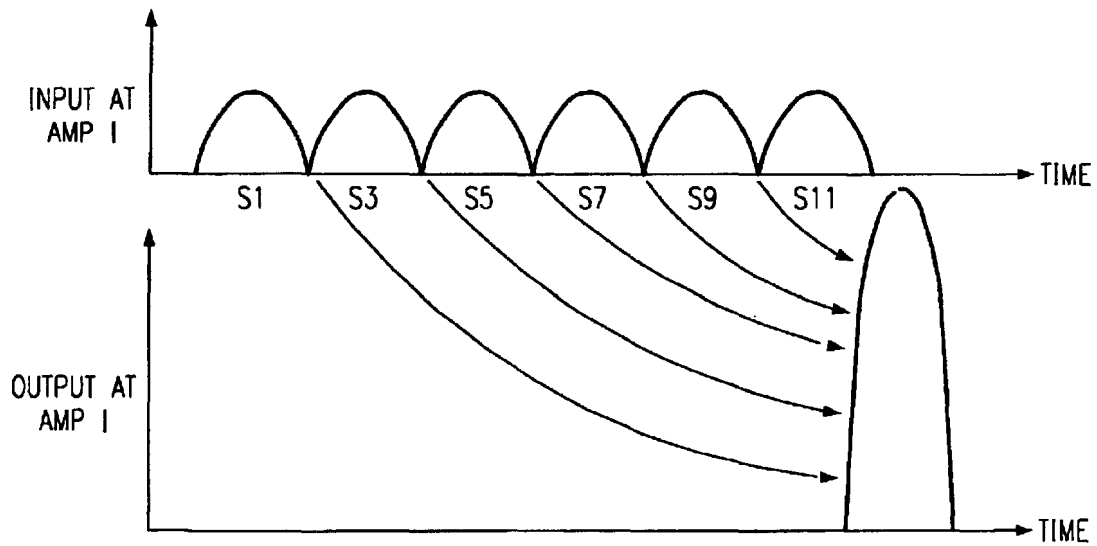


FIG. 9

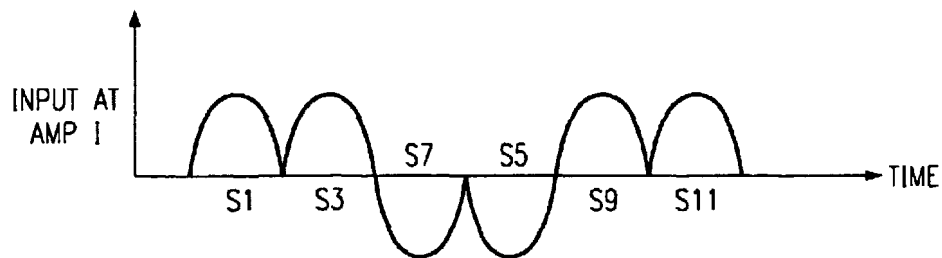


FIG. 10

**MULTI-TAP, DIGITAL-PULSE-DRIVEN MIXER**

This application claims the priority under 35 U.S.C. 119(e)(1) of now abandoned U.S. provisional application No. 60/195,926 filed on Apr. 10, 2000.

**FIELD OF THE INVENTION**

The invention relates generally to frequency channel communications and, more particularly, to mixers for down-converting the frequency of a received communication signal.

**BACKGROUND OF THE INVENTION**

Among conventional RF-to-IF (radio frequency-to-intermediate frequency) mixers, zero-IF implementations have the inherent problem of LO (local oscillator) leakage through the mixer to the RF input, which then gets down-converted inside the mixer. One solution currently in discussion to solve this problem is to use a sub-harmonic pumped mixer for the down conversion. Such a mixer either requires very high LO drive currents or suffers from undesirably high noise figures. The sub-harmonic pumped mixer also generates the needed RF frequency internally, so there is still a LO leakage problem with this architecture. For low IF implementations, the realization of 90 degree phase splitters is one of the biggest challenges.

It is therefore desirable to provide a mixer that avoids the aforementioned disadvantages of conventional approaches.

**SUMMARY OF THE INVENTION**

The present invention provides a multi-tap, digital-pulse-driven mixer which advantageously avoids LO leakage by shifting the LO frequency out of the receive frequency band, and which advantageously realizes low noise figures by the use of digital pulses as mixer drive signals.

**BRIEF DESCRIPTION OF THE DRAWINGS**

FIG. 1 diagrammatically illustrates an exemplary mixer embodiment according to the invention.

FIG. 2 diagrammatically illustrates exemplary embodiments of the switches and anti-aliasing filters of FIG. 1.

FIG. 3 graphically illustrates an exemplary timing relationship between the RF signal of FIG. 1 and the sampling pulse signal of FIG. 1.

FIG. 4 diagrammatically illustrates an exemplary embodiment of the delay element section of FIG. 1.

FIG. 5 graphically illustrates examples of various signals from FIG. 4, their mutual timing relationships, and their respective timing relationships relative to the RF signal of FIG. 1.

FIG. 6 illustrates in tabular format exemplary operations of the router of FIG. 1.

FIG. 7 illustrates exemplary operations which can be performed by the embodiments of FIGS. 1-6.

FIG. 8 illustrates exemplary operations which can be performed by the embodiments of FIGS. 1-6.

FIGS. 9 and 10 graphically illustrate exemplary signals of FIG. 1.

**DETAILED DESCRIPTION**

If a mixer, such as an RF-to-IF mixer, is driven using a digital pulse with rise and fall times that are small compared

to the pulse width, the needed voltage swing can be reduced. For a resistive mixer, which is equivalent to a sampling switch, the needed voltage swing above  $V_{th}$  is determined by the  $g_m$  saturation at low  $V_{ds}$ , because the maximum signal swing at that point in, for example a wireless system, is limited to 50 mV. Therefore 150 to 200 mV will be sufficient. The voltage swing needed below  $V_{th}$  is determined by the off current needed and will be around 300-400 mV. Therefore a total voltage swing of 500-600 mV will be sufficient. This voltage swing can be realized, for example, by local power regulation of the driving inverter. When comparing that situation to, for example, an analog mixer drive circuit with a sinusoidal drive waveform and wherein the needed overdrive voltage is small, one can obtain an equivalent analog voltage amplitude by calculating waveforms with equal voltage derivatives at zero crossing points:

$$dV_{anal}/dt = dV_{dg}/dt \tag{Equation 1}$$

$$d(V_{pp}/2 * \sin(\omega t)) = V_S/t_{rr} \tag{Equation 2}$$

$$V_{pp} = V_S * 1/\pi * (T_{RF}/t_{rr}) \tag{Equation 3}$$

where  $V_S$  is the digital voltage swing and  $t_{rr}$  is the digital transition (rise/fall) time.

Equation 3 shows that a factor of  $1/\pi * (T_{RF}/t_{rr})$  is gained with respect to voltage swing. For typical inverter delays of 20 psec for a conventional Texas Instruments deep-submicron CMOS process with  $L_g=0.13$  micrometers, the gain with a digital-pulse-driven mixer can be around a factor of 10 compared to an analog implementation.

The current consumption needed by a digital drive circuit can also be calculated. The mean current consumption for one digital pulse with a repetition rate  $T_{rep}$ , is given by:

$$I_{mean} = 2 * V_S * (C_{load} + C_{par} + C_{inv}) / T_{rep} \tag{Equation 4}$$

where  $C_{load}$  is the capacitance of the sampling switch,  $C_{par}$  is the parasitic capacitance of the wiring and  $C_{inv}$  is the output capacitance of, for example, a driving inverter. It is important to notice that the current consumption is independent of  $t_{rr}$ , while the noise figure of the circuit goes down with  $t_{rr}$ . The size of  $C_{load}$  is determined by the needed on resistance  $R_{on}$  of the sampling switch, which should be a factor of 10 lower than the input impedance of the first IF amplifier. To meet typical noise floor requirements with an exemplary LNA gain of 20 dB, an input impedance around 500 Ohm or a sample switch  $g_m$  of around 50 Ohm is required. With a typical  $g_m$  of 3 mS/um, a 50 um wide transistor is needed, which has an input capacitance of 40 fF. The typical output capacitance of an inverter is very similar, and the interconnection parasitic can be kept below 5 fF with suitable attention to the layout. This gives total current consumption of 0.25 mA.

FIG. 1 diagrammatically illustrates an exemplary embodiment of a mixer according to the present invention for downconverting a communication signal from RF (radio frequency) to IF (intermediate frequency). The exemplary embodiment of FIG. 1 is a digital-pulse-driven mixer which can, accordingly, realize one or more of the aforementioned advantageous characteristics associated with a digital-pulse-driven design. In FIG. 1, an RF communication signal input **22** is applied to a low noise amplifier (LNA) **18** whose output **23** is in turn applied to a plurality of sampling switches **19**. In response to a plurality of digital control signals **16**, the sampling switches at **19** sample the amplified RF signal **23**. The switches **19** output the sampled RF signal at **20** to anti-aliasing filters **21**, which produce the IF signal.

A local oscillator **11** produces a synthesized frequency signal **12** having a frequency  $F_{LO}$ . This local oscillator signal

**12** is input to a digital pulse generator **13** which produces in response thereto a sampling pulse signal SPS which is in turn input to a section of delay elements **15**. A router **17** is coupled to receive signals **14** from the outputs of the respective delay elements at **15**, and the router **17** also receives the sampling pulse signal SPS. The router **17** suitably routes the signals **14** and the sampling pulse signal SPS to drive the various digital control signals **16** and thereby control the sampling switches **19** as desired. The router **17** and switches **19** thus provide a sampler for sampling the RF signal **23**.

FIG. 2 diagrammatically illustrates exemplary embodiments of selected portions of the mixer of FIG. 1. In the example of FIG. 2, the switches **19** are provided as CMOS pass gates controlled by the digital signals **16** produced by the router **17**. The exemplary embodiment of FIG. 2 includes  $n$  switches  $S1-Sn$ , where  $n=M \times 4$  and  $M$  is an integer. The switches **19** are partitioned into  $M$  groups of 4 switches, the switches  $S1-S4$  being exemplary of one such group. As shown in FIG. 2, switch  $S1$  samples the RF input signal **23** at a phase of  $0^\circ$ , switch  $S2$  samples at a phase of  $90^\circ$ , switch  $S3$  samples at  $180^\circ$ , and switch  $S4$  samples at  $270^\circ$ . Similarly, switches  $S5, S9, \dots, Sn-3$  sample at  $0^\circ$ , switches  $S6, S10, \dots, Sn-2$  sample at  $90^\circ$ , switches  $S7, S11, \dots, Sn-1$  sample at  $180^\circ$ , and switches  $S8, S12, \dots, Sn$  sample at  $270^\circ$ . The sampled phases are input to appropriate anti-aliasing filters **21** which recombine the sampled phases. In the example of FIG. 2, the anti-aliasing filters **21** are conventional third-order low-pass filters, one of which includes an in-phase IF amplifier I that receives phases  $0^\circ$  and  $180^\circ$ , and the other of which includes a quadrature IF amplifier Q that receives phases  $90^\circ$  and  $270^\circ$ . The outputs of the filters **21** can be applied to, for example, a conventional  $\Sigma\Delta$  multi-bit A/D converter (not shown).

Referring also to FIG. 1,  $n-1$  of the  $n$  digital control signals **16** are provided as delayed versions of a pulse (or pulses) of the sampling pulse signal SPS, and one of the control signals **16** is the pulse (or one of the pulses) from which the delayed versions are produced. For example, if switch  $S1$  is controlled by a given SPS pulse, then switches  $S2-Sn$  can be driven by respective delayed versions of that SPS pulse. If each of the four phases is sampled during each cycle of the RF input signal **23**, then a new SPS pulse will be needed approximately every  $M (=n/4)$  cycles of the signal **23**.

Advantageously according to the invention, the SPS pulses have a pulse width which is approximately equal to but slightly larger than the half period of the RF input signal, as illustrated generally in FIG. 3. The FIG. 3 relationship between the SPS pulse width and the half period of the RF input signal can advantageously reduce the noise figure of the mixer, because the switching point of at least some of the pulses which control the sampling operations of switches  $S1-Sn$  (see also FIG. 2) can be made exactly aligned with the zero crossings of the RF signal **23**, which allows implementation of coherent detection. As one example, the SPS pulse width can be  $[(n+1)/n] \times (\text{half period of RF input signal})$ . In this example, the relationship of the frequency  $F_{LO}$  of the local oscillator output **12** (see also FIG. 1) to the frequency  $F_{RF}$  of the RF input signal should be:  $F_{LO} = F_{RF} \times [n/(n+1)]$ . The digital pulse generator **13** of FIG. 1 can then utilize well-known conventional techniques to produce the sampling pulse signal SPS having a pulse duration of  $[(n+1)/n] \times (\text{half period of RF input signal})$  and such that the SPS pulse is repeated every  $M$  cycles of the local oscillator output **12**.

Due to the above-described exemplary relationship between  $F_{LO}$  and  $F_{RF}$ , the length of each cycle of the local

oscillator output **12** will be  $[1+(1/n)] \times (\text{period of the RF input signal})$ . Recalling that the spacing between SPS pulses is  $M$  cycles of the local oscillator output **12**, and recalling that  $M=n/4$ , the timing relationship of the  $(j+1)$ th SPS pulse with respect to the RF input signal will be delayed by  $1/4$  of a cycle of the RF input signal when compared to the timing relationship of the immediately preceding  $(j)$ th SPS pulse with respect to the RF input signal. This  $1/4$  of a cycle delay is due to the fact that the local oscillator signal **12** "loses"  $(1/n)$ th of a cycle (relative to the RF signal **23**) during each of the  $M=n/4$  cycles between SPS pulses, and

$$\frac{1}{n} \times \frac{n}{4} = \frac{1}{4}.$$

This delay between adjacent SPS pulses can be compensated for in the design of the delay elements **15** and the router **17** of FIG. 1, as described in detail below.

FIG. 4 diagrammatically illustrates an exemplary embodiment of the delay element section **15** of FIG. 1. The embodiment of FIG. 4 includes a plurality of delay elements  $DE1-DEn-1$  and DEC connected in series to form a delay chain. In some embodiments, each of the illustrated delay elements provides a delay of  $1/4$  cycle of the RF input signal **23**. Referring also to FIGS. 1 and 2, the router **17** can route SPS to control switch  $S1$ , and can also route the outputs of delay elements  $DE1-DEn-1$  to respectively control switches  $S2-Sn$ . Because each of the delay elements delays the input SPS pulse by  $1/4$  of a cycle of the RF input signal, the SPS pulse and the respective  $1/4$  cycle delayed versions thereof can control switches  $S1-Sn$  to sample at the appropriate phases of the RF input signal.

For example, the SPS pulse can be used to control switch  $S1$  to sample at  $0^\circ$ , the output of delay element  $DE1$  can be used to control switch  $S2$  to sample at  $90^\circ$ , the output of delay element  $DE2$  can be used to control switch  $S3$  to sample at  $180^\circ$ , and the output of delay element  $DE3$  can be used to control switch  $S4$  to sample at  $270^\circ$ . The delay element  $DE4$  can be used to control the next switch  $S5$  (not illustrated in FIG. 2) to sample at  $0^\circ$  of the next cycle of the RF input signal **23**, and so on until delay element  $DEn-1$  controls switch  $Sn$  to sample at  $270^\circ$  of the  $M$ th cycle of signal **23**. This exemplary operation is illustrated generally in FIG. 5.

As shown in FIG. 5, the SPS pulse **51** provides for sampling at  $0^\circ$  of cycle **1** of the RF signal **23** and the sampling continues at  $90^\circ$  phase increments through the sampling at  $270^\circ$  of cycle  $M$  by delay element  $DEn-1$ . As mentioned above, however, after  $M$  cycles of the local oscillator output **12**, the timing relationship of the next SPS pulse **52** with respect to the RF input signal **23** will be delayed by  $1/4$  cycle ( $90^\circ$  phase) as compared to the timing relationship of the SPS pulse **51** with respect to the RF input signal **23**. Thus, as illustrated in FIG. 5, the SPS pulse **52** will not be available to sample at  $0^\circ$  of cycle  $M+1$  of the RF input signal, but rather will be available  $1/4$  of a cycle later to sample at  $90^\circ$  of cycle  $M+1$ . Accordingly, the router **17** of FIG. 1 can route the SPS pulse **52** to switch  $S2$  of FIG. 2 for sampling at  $90^\circ$  of cycle  $M+1$ . The sampling at  $0^\circ$  of cycle  $M+1$  is controlled by the pulse output from the compensating delay element DEC, which the router **17** routes to control the switch  $S1$  of FIG. 2. The output of  $DE1$  is routed to switch  $S3$  to sample at  $180^\circ$  of cycle  $M+1$ , the output of  $DE2$  is routed to switch  $S4$  to sample at  $270^\circ$  of cycle  $M+1$ , and so on as illustrated in FIG. 5.

FIG. 6 illustrates in tabular format exemplary operations which can be performed by the router **17** of FIG. 1 to control



5

the sampling switches at **19** in FIG. 2. The example of FIG. **6** is for  $n=16$  switches partitioned into  $M=4$  groups of 4 switches each, each group of 4 switches operable for sampling the desired 4 phases of an associated cycle of the RF input signal. Also in the FIG. **6** example,  $F_{LO}=F_{RF} \times [n/(n+1)] = F_{RF} \times (16/17)$ . As shown in FIG. **6**, for a given cycle  $K$  of the RF input signal, the SPS pulse (e.g., **51** in FIG. **5**) is used to control switch **S1** to sample at  $0^\circ$ , and the respective delay elements **DE1–DE15** are used to control the respective switches **S2–S16** to sample as shown in cycles  $K$  through  $K+3$ . In cycle  $K+4$ , the output of **DEC** is used to control switch **S1** to sample at  $0^\circ$ , the SPS pulse (e.g., **52** in FIG. **5**) is used to control switch **S2** to sample at  $90^\circ$ , and the respective outputs of **DE1–DE14** are used to control the respective sampling operations of the switches **S3–S16** in the remainder of cycle  $K+4$  and in cycles  $K+5$  through  $K+7$ .

In cycle  $K+8$ , the output of **DE15** is used to control switch **S1** to sample at  $0^\circ$ , the output of **DEC** is used to control switch **S2** for sampling at  $90^\circ$ , and the SPS pulse is used to control switch **S3** for sampling at  $180^\circ$ . The output of **DE1** is used to control **S4** for sampling at  $270^\circ$  during cycle  $K+8$ , and the respective outputs of **DE2–DE13** are used to control the sampling operations of the respective switches **S5–S16** in cycles  $K+9$  through  $K+11$ . In cycle  $K+12$ , the output of **DE14** drives switch **S1** to sample at  $0^\circ$ , the output of **DE15** drives switch **S2** to sample at  $90^\circ$ , the output of **DEC** drives switch **S3** to sample at  $180^\circ$ , and the SPS pulse drives **S4** to sample at  $270^\circ$ . The respective outputs of **DE1–DE12** are used to control the respective sampling operations of switches **S5–S16** in cycles  $K+13$  through  $K+15$ .

In the next cycle of the RF input signal, namely cycle  $K+16$ , the SPS pulse will be back in proper position relative to the RF input signal for controlling switch **S1**, **S5**, **S9** or **S13** to sample at  $0^\circ$ . This is because, in this example, after 16 cycles ( $K$  through  $K+15$ ) of the RF input signal, the SPS pulse now “lags” the RF input signal by  $16 \times \frac{1}{16} = 1$  cycle, and is thus back in its “original” phase (i.e., its cycle  $K$  phase) relative to the RF signal. Accordingly, after cycle  $K+15$ , the operations in FIG. **6** can, for example, return to cycle  $K$  and repeat (in which case the SPS pulse would control switch **S1** again).

The router **17** can be readily implemented, for example, utilizing a passive pass gate design including a matrix of CMOS pass gates controlled by bits in a plurality of  $n$ -bit registers. In the example of FIG. **6**, a total of four  $n$ -bit registers can be used, each register corresponding to a respective one of the four routing schemes shown in

FIG. **6**. The registers can be sequentially enabled (one every 4<sup>th</sup> RF cycle) in the cyclic pattern illustrated in FIG. **6**.

FIG. **7** illustrates exemplary operations which can be performed by the embodiments illustrated in FIGS. **1–6**. At **71**, the local oscillator frequency  $F_{LO}$  is set to be less than the frequency  $F_{RF}$  of the RF input signal, and a sample switch index  $i$  is set to **1**. At **72**, the sampling pulse signal SPS is produced from the local oscillator. At **73**, the  $j$ th SPS pulse is selected as the current sample pulse, and is applied to switch  $S_i$  at **74**. For example, the  $j$ th SPS pulse can be applied to switch **S1** in order to sample at  $0^\circ$ . Thereafter, if it is determined at **75** that switch  $S_n$  has not yet been operated, then the next switch is selected at **70** by incrementing the switch index  $i$ . Thereafter at **76**, a new sample pulse is produced in response to the current sample pulse, for example by producing a delayed version of the SPS pulse that was selected at **73**. At **77**, the new sample pulse is selected as the current sample pulse, and the current sample pulse is applied to switch  $S_i$  at **74**. The above-described

6

operations at **70** and **74–77** are repeated until it is determined at **75** that all  $n$  switches have been operated.

When it is determined at **75** that all  $n$  switches have been operated, at **79** the sampling switch index  $i$  is again set equal to **1**, and the SPS pulse index  $j$  is incremented. It is thereafter determined at **78** whether the  $j$ th SPS pulse is in phase for the assigned sampling operation of switch  $S_i$ . If not, then the operations described above and illustrated at **76**, **77** and **74** are sequentially executed in that order. Thereafter, the sampling switch index  $i$  is incremented at **80**, after which it is determined at **78** whether the  $j$ th SPS pulse is in the appropriate phase for controlling the assigned sampling operation of switch  $S_i$ . If not, then the aforementioned sequence of operations **76**, **77**, **74**, **80** and **78** are repeated. However, if the  $j$ th SPS pulse is determined at **78** to be in the appropriate phase for controlling the assigned sampling operation of switch  $S_i$ , then the  $j$ th SPS pulse is at **73** selected to be the current sample pulse. Thereafter, operations beginning at **74** are repeated again as described above.

FIG. **8** illustrates exemplary operations which can be performed by the embodiments of FIGS. **1–6**. After generation of an SPS pulse at **81**, that pulse and delayed versions thereof are used at **82** to sample desired phases in adjacent cycles of the RF signal. The sampled phases are recombined at **83** to produce the desired downconverted signal.

As will be evident to workers in the art, the embodiments of FIGS. **1–8** can be used to realize a zero-IF or near-zero-IF receiver architecture wherein the frequency of the local oscillator is advantageously shifted away from the frequency of the RF input signal by a factor such as  $n/(n+1)$ . For example, in the case of a Bluetooth receiver with  $n=16$ , the oscillator frequency is 2.25 GHz for an RF input frequency of 2.4 GHz, and the oscillator frequency is 2.34 GHz for an RF input frequency of 2.5 GHz. Thus, the frequency of the local oscillator lies outside of the Bluetooth frequency band, which insures that any leakage from the local oscillator is suppressed by the Bluetooth antenna filter, and also insures that no other channel is folded into the downconverted signal. The local oscillator can therefore be advantageously integrated without the leakage problem of conventional arrangements. Also, the delay elements can be realized, for example, by a suitable inverter chain, which advantageously requires a much smaller silicon area than conventional polyphase networks. Furthermore, because all desired phases of each cycle of the RF input signal are sampled and recombined in the IF amplifier, there is no signal loss as compared to a conventional sub-sampling scheme. This is illustrated by FIG. **9**, which shows the in-phase path.

In some embodiments, the router **17** can control the switches **19** to generate an SC (switched capacitor) filter function during the phase sampling operations. In this manner, undesired interferers can be advantageously attenuated during the sampled-phase recombination operations of the IF amplifiers. An example of this is illustrated by FIG. **10**, wherein the switch activation sequence is modified as shown (**S5** and **S7** are reversed with respect to the sequence of FIG. **9**) to support a desired SC filter function.

Although exemplary embodiments of the invention are described above in detail, this does not limit the scope of the invention, which can be practiced in a variety of embodiments.

What is claimed is:

1. A method of downconverting a first communication signal at a first frequency into a second communication signal at a second frequency that is lower than the first frequency, comprising:

providing an oscillator signal having a third frequency that is lower than the first frequency;

producing in response to the oscillator signal a sampling pulse signal having digital pulses for use in sampling the first communication signal, wherein adjacent pulses of the sampling pulse signal are separated by an amount of time that corresponds to a predetermined number of cycles of the first communication signal and wherein said amount of time is greater than an amount of time required for completion of said predetermined number of cycles of the first communication signal;

using the pulses of the sampling pulse signal to sample selected phases of the first communication signal, including using a first pulse of the sampling pulse signal to sample a first phase of a first cycle of the first communication signal and using a second pulse of the sampling pulse signal to sample a second phase of a second cycle of the first communication signal in which said first and second pulses are adjacent one another in the sampling pulse signal, wherein said second phase is a different phase than said first phase, and wherein said second cycle follows said first cycle by a number of cycles of the first communication signal equal to said predetermined number; and

using the sampled phases to produce the second communication signal.

2. The method of claim 1, wherein the first communication signal is an RF communication signal.

3. The method of claim 1, wherein said first-mentioned using step includes using a delayed version of said first pulse to sample said first phase of said second cycle.

4. The method of claim 1, wherein said first-mentioned using step includes using a third pulse of the sampling pulse signal to sample said first phase of a third cycle of the first communication signal, and wherein said third cycle follows said second cycle.

5. The method of claim 4, wherein said third cycle follows said first cycle by a number of cycles of the first communication signal that is a multiple of said predetermined number.

6. The method of claim 5, wherein said first-mentioned using step includes using a delayed version of said first pulse to sample said first phase of said second cycle.

7. The method of claim 4, wherein said first-mentioned using step includes using a delayed version of said first pulse to sample said first phase of said second cycle.

8. The method of claim 1, wherein said step of using pulses includes using a first pulse of the sampling pulse signal and a plurality of delayed versions of said first pulse to sample selected phases of the first communication signal.

9. A method of downconverting a first communication signal at a first frequency into a second communication signal at a second frequency that is lower than the first frequency, comprising:

sampling a plurality of phases of each of at least two consecutive cycles of the first communication signal, wherein said sampling step includes normally activating a plurality of sampling switches in a first temporal order to sample said plurality of phases, and providing said filter function by activating said plurality of switches in a second temporal order that differs from said first temporal order; and

combining the sampled phases to provide a filter function and produce the second communication signal.

10. An apparatus for downconverting a first communication signal at a first frequency into a second communication signal at a second frequency that is lower than the first frequency, comprising:

an oscillator for generating a signal having a third frequency that is lower than the first frequency;

circuitry for producing in response to the oscillator signal a sampling pulse signal having digital pulses for use in sampling the first communication signal, wherein adjacent pulses of the sampling pulse signal are separated by an amount of time that corresponds to a predetermined number of cycles of the first communication signal and wherein said amount of time is greater than an amount of time required for completion of said predetermined number of cycles of the first communication signal;

circuitry for using the pulses of the sampling pulse signal to sample selected phases of the first communication signal, including using a first pulse of the sampling pulse signal to sample a first phase of a first cycle of the first communication signal and using a second pulse of the sampling pulse signal to sample a second phase of a second cycle of the first communication signal in which said first and second pulses are adjacent one another in the sampling pulse signal, wherein said second phase is a different phase than said first phase, and wherein said second cycle follows said first cycle by a number of cycles of the first communication signal equal to said predetermined number; and

circuitry for using the sampled phases to produce the second communication signal.

11. The apparatus of claim 10, wherein said circuitry for using the pulses further includes circuitry for using a delayed version of said first pulse to sample said first phase of said second cycle.

12. The apparatus of claim 10, wherein said circuitry for using the pulses further includes circuitry for using a third pulse of the sampling pulse signal to sample said first phase of a third cycle of the first communication signal, and wherein said third cycle follows said second cycle.

13. The apparatus of claim 12, wherein said third cycle follows said first cycle by a number of cycles of the first communication signal that is a multiple of said predetermined number.

14. The apparatus of claim 13, wherein said circuitry for using the pulses further includes circuitry for using a delayed version of said first pulse to sample said first phase of said second cycle.

15. The apparatus of claim 12, wherein said circuitry for using the pulses further includes circuitry for using a delayed version of said first pulse to sample said first phase of said second cycle.

16. An apparatus for downconverting a first communication signal at a first frequency into a second communication signal at a second frequency that is lower than the first frequency, comprising:

circuitry for sampling a plurality of phases of each of at least two consecutive cycles of the first communication signal, wherein said sampling includes normally activating a plurality of sampling switches in a first temporal order to sample said plurality of phases, and providing a filter function by activating a plurality of switches in a second temporal order that differs from said first temporal order; and

circuitry for combining the sampled phases to provide said filter function and produce the second communication signal.

17. The apparatus of claim 10, wherein said circuitry for producing comprises a sampler, said sampler including a plurality of sampling switches coupled to an input for sampling the first communication signal.

18. The apparatus of claim 10, wherein said circuitry for producing comprises a sampler operable for sampling a plurality of phases of all cycles of the first communication signal.

19. The apparatus of claim 17, including a digital pulse generator coupled to said sampler for producing a sampling pulse signal having a plurality of digital pulses, each of said pulses having a pulse width that is approximately equal to but wider than a half period of the first communication signal, said sampler responsive to said sampling pulse signal for sampling the first communication signal.

20. The apparatus of claim 18, including a digital pulse generator coupled to said sampler for producing a sampling pulse signal having a plurality of digital pulses, each of said pulses having a pulse width that is approximately equal to but wider than a half period of the first communication signal, said sampler responsive to said sampling pulse signal for sampling the first communication signal.

21. The apparatus of claim 19, wherein said sampler has an input for receiving one of said digital pulses and a plurality of delayed versions of said one digital pulse, said sampler responsive to said one digital pulse for sampling one of the phases of said consecutive cycles, and said sampler responsive to said delayed versions of said one pulse for sampling other phases of said consecutive cycles.

22. The apparatus of claim 20, wherein said sampler has an input for receiving one of said digital pulses and a plurality of delayed versions of said one digital pulse, said sampler responsive to said one digital pulse for sampling one of the phases of said consecutive cycles, and said sampler responsive to said delayed versions of said one pulse for sampling other phases of said consecutive cycles.

23. The apparatus of claim 21, further including a delay element structure coupled to said digital pulse generator and said sampler for producing the delayed versions of said one pulse and providing the delayed versions to said sampler input.

24. The apparatus of claim 22, further including a delay element structure coupled to said digital pulse generator and said sampler for producing the delayed versions of said one pulse and providing the delayed versions to said sampler input.

25. The apparatus of claim 21, wherein said sampler includes a plurality of sampling switches coupled to said first-mentioned input and to said sampler input for respectively sampling phases of said consecutive cycles of the first communication signal in response to said one pulse and said delayed versions of said one pulse.

26. The apparatus of claim 22, wherein said sampler includes a plurality of sampling switches coupled to said first-mentioned input and to said sampler input for respectively sampling phases of said consecutive cycles of the first communication signal in response to said one pulse and said delayed versions of said one pulse.

27. The apparatus of claim 10, wherein the first communication signal is an RF communication signal.

28. The apparatus of claim 10, wherein the circuitry for using the sampled phases includes filters respectively for receiving selected ones of the sampled phases.

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