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(12) **United States Patent**  
**Sorrells et al.**

(10) **Patent No.:** **US 8,571,135 B2**  
(45) **Date of Patent:** **\*Oct. 29, 2013**

(54) **METHOD, SYSTEM AND APPARATUS FOR BALANCED FREQUENCY UP-CONVERSION OF A BASEBAND SIGNAL**

7, 1999, provisional application No. 60/171,349, filed on Dec. 21, 1999, provisional application No. 60/177,702, filed on Jan. 24, 2000, provisional application No. 60/180,667, filed on Feb. 7, 2000, provisional application No. 60/171,496, filed on Dec. 22, 1999.

(75) Inventors: **David F. Sorrells**, Middleburg, FL (US); **Michael J. Bultman**, Jacksonville, FL (US); **Robert W. Cook**, Switzerland, FL (US); **Richard C Looke**, Jacksonville, FL (US); **Charley D. Moses, Jr.**, DeBary, FL (US); **Gregory S. Rawlins**, Chuluota, FL (US); **Michael W. Rawlins**, Lake Mary, FL (US)

(51) **Int. Cl.**  
**H04L 27/04** (2006.01)  
**H04L 27/12** (2006.01)  
**H04L 27/20** (2006.01)

(73) Assignee: **ParkerVision, Inc.**, Jacksonville, FL (US)

(52) **U.S. Cl.**  
USPC ..... **375/295**; 375/298; 375/259; 375/256; 455/76; 455/91

(\* ) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(58) **Field of Classification Search**  
USPC ..... 375/295, 298, 259, 256, 296, 309-312, 375/268; 455/76, 91, 118, 323, 313  
See application file for complete search history.

This patent is subject to a terminal disclaimer.

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(22) Filed: **Dec. 12, 2011**

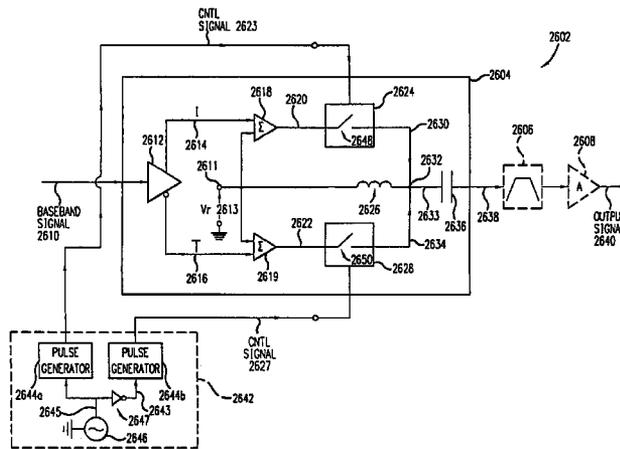
(65) **Prior Publication Data**

US 2012/0114078 A1 May 10, 2012

**Related U.S. Application Data**

(63) Continuation of application No. 12/823,055, filed on Jun. 24, 2010, now Pat. No. 8,077,797, which is a continuation of application No. 11/015,653, filed on Dec. 20, 2004, now Pat. No. 7,773,688, which is a continuation of application No. 09/525,615, filed on Mar. 14, 2000, now Pat. No. 6,853,690.

(60) Provisional application No. 60/177,381, filed on Jan. 24, 2000, provisional application No. 60/171,502, filed on Dec. 22, 1999, provisional application No. 60/177,705, filed on Jan. 24, 2000, provisional application No. 60/129,839, filed on Apr. 16, 1999, provisional application No. 60/158,047, filed on Oct.



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*Primary Examiner* — Phuong Phu  
 (74) *Attorney, Agent, or Firm* — Workman Nydegger

(57) **ABSTRACT**

A balanced transmitter up-converts a baseband signal directly from baseband-to-RF. The up-conversion process is sufficiently linear that no IF processing is required, even in communications applications that have stringent requirements on spectral growth. In operation, the balanced modulator subharmonically samples the baseband signal in a balanced and differential manner, resulting in harmonically rich signal. The harmonically rich signal contains multiple harmonic images that repeat at multiples of the sampling frequency, where each harmonic contains the necessary information to reconstruct the baseband signal. The differential sampling is performed according to a first and second control signals that are phase shifted with respect to each other. In embodiments of the invention, the control signals have pulse widths (or apertures) that operate to improve energy transfer to a desired harmonic in the harmonically rich signal. A bandpass filter can then be utilized to select the desired harmonic of interest from the harmonically rich signal. The sampling modules that perform the sampling can be configured in either a series or a shunt configuration. In embodiments of the invention, DC offset voltages are minimized between the sampling modules to minimize or prevent carrier insertion into the harmonic images.

**24 Claims, 144 Drawing Sheets**

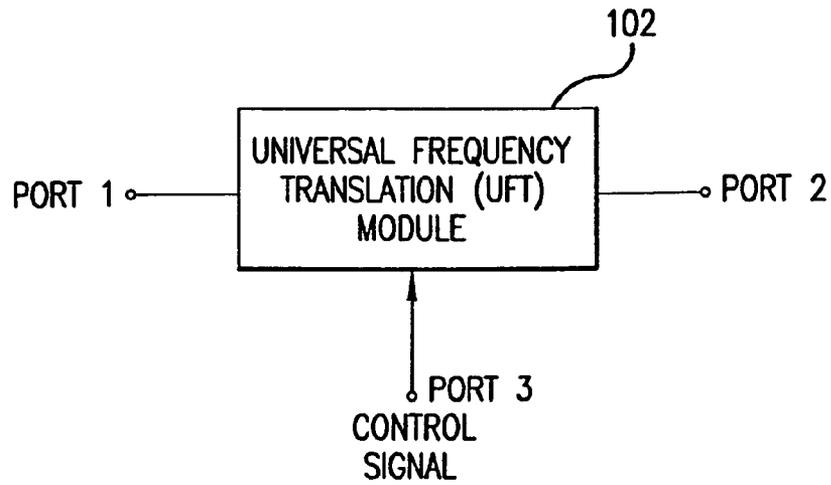


FIG. 1A

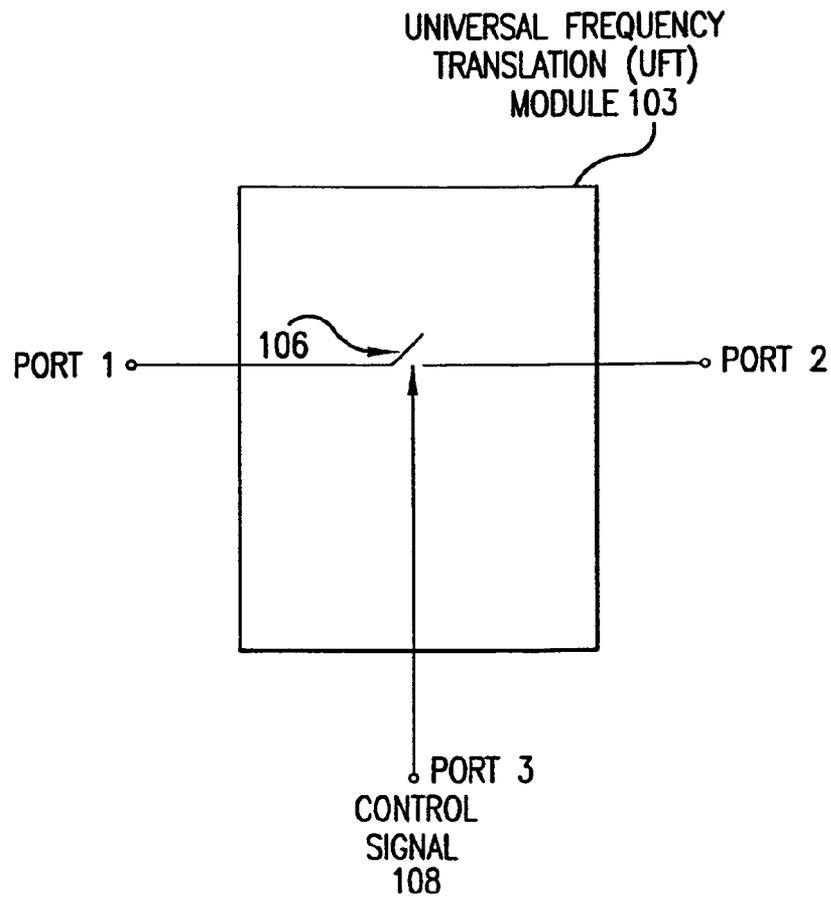


FIG. 1B

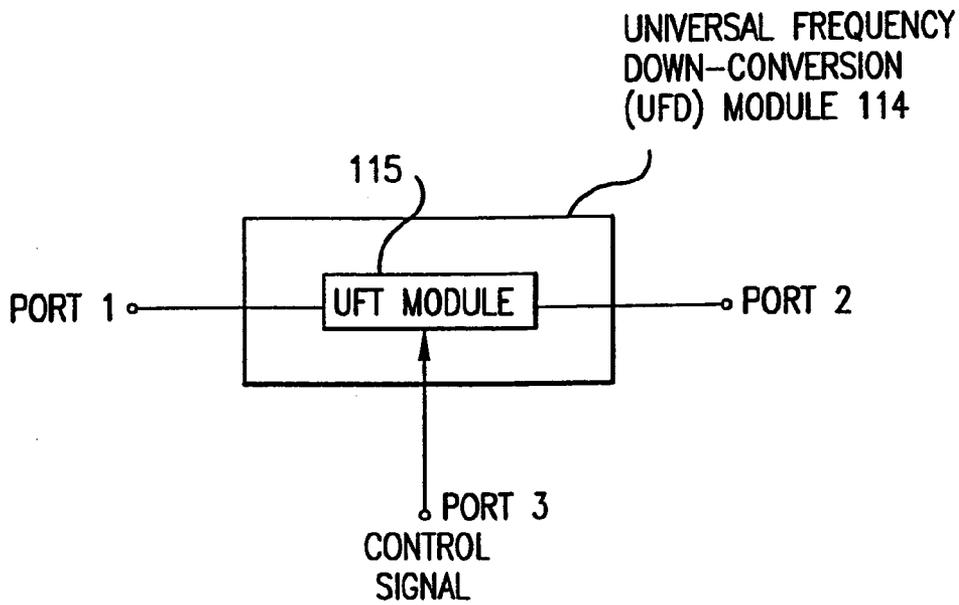


FIG. 1C

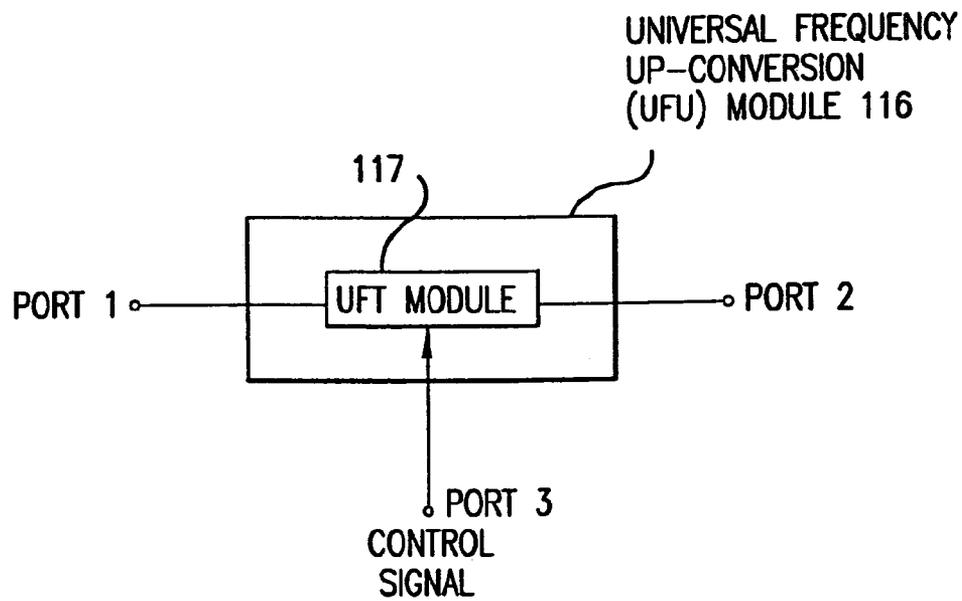
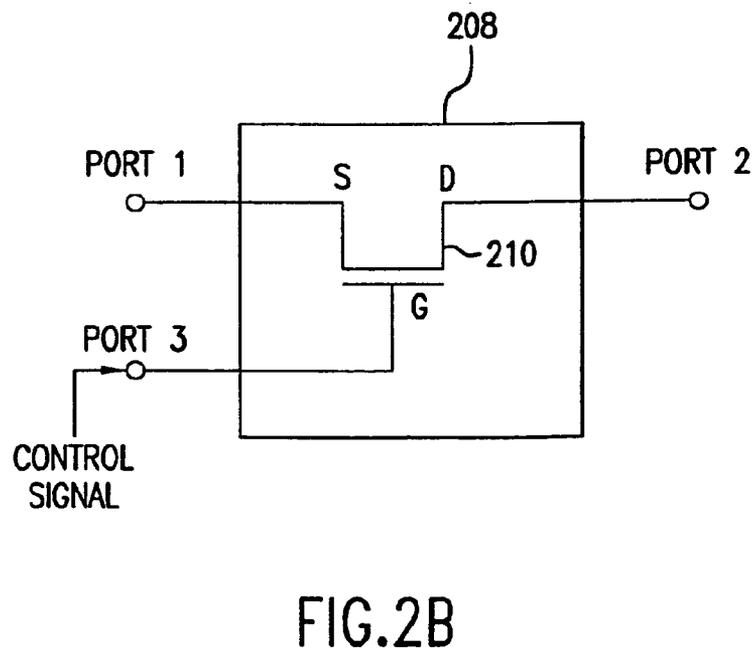
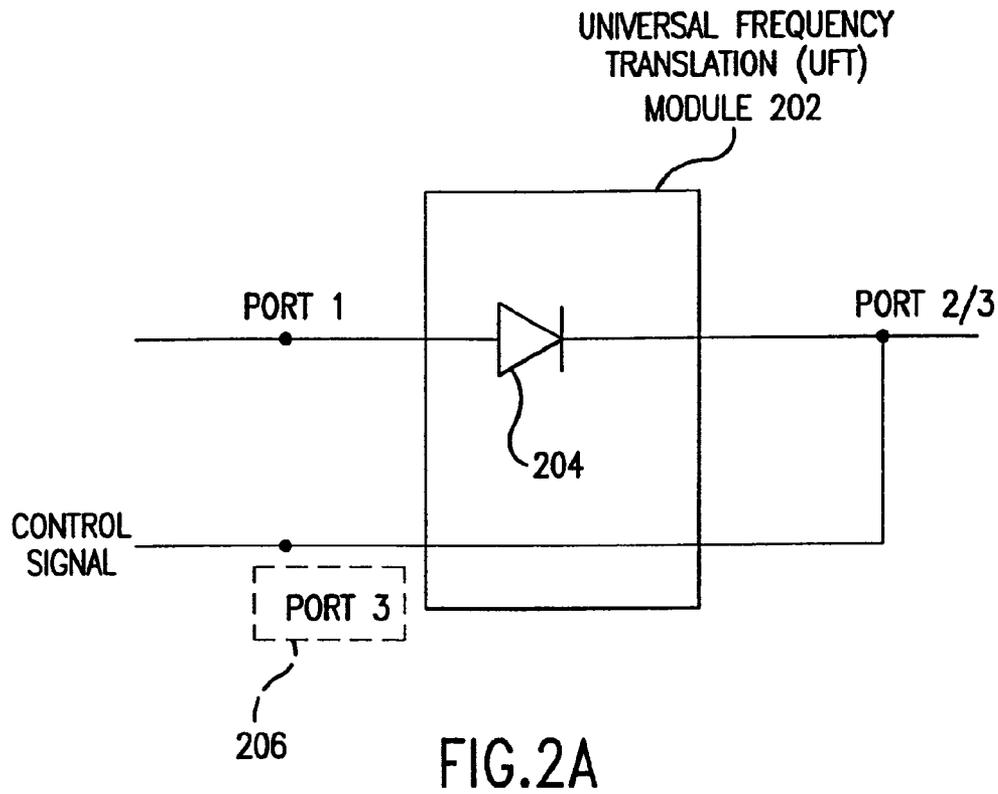


FIG. 1D



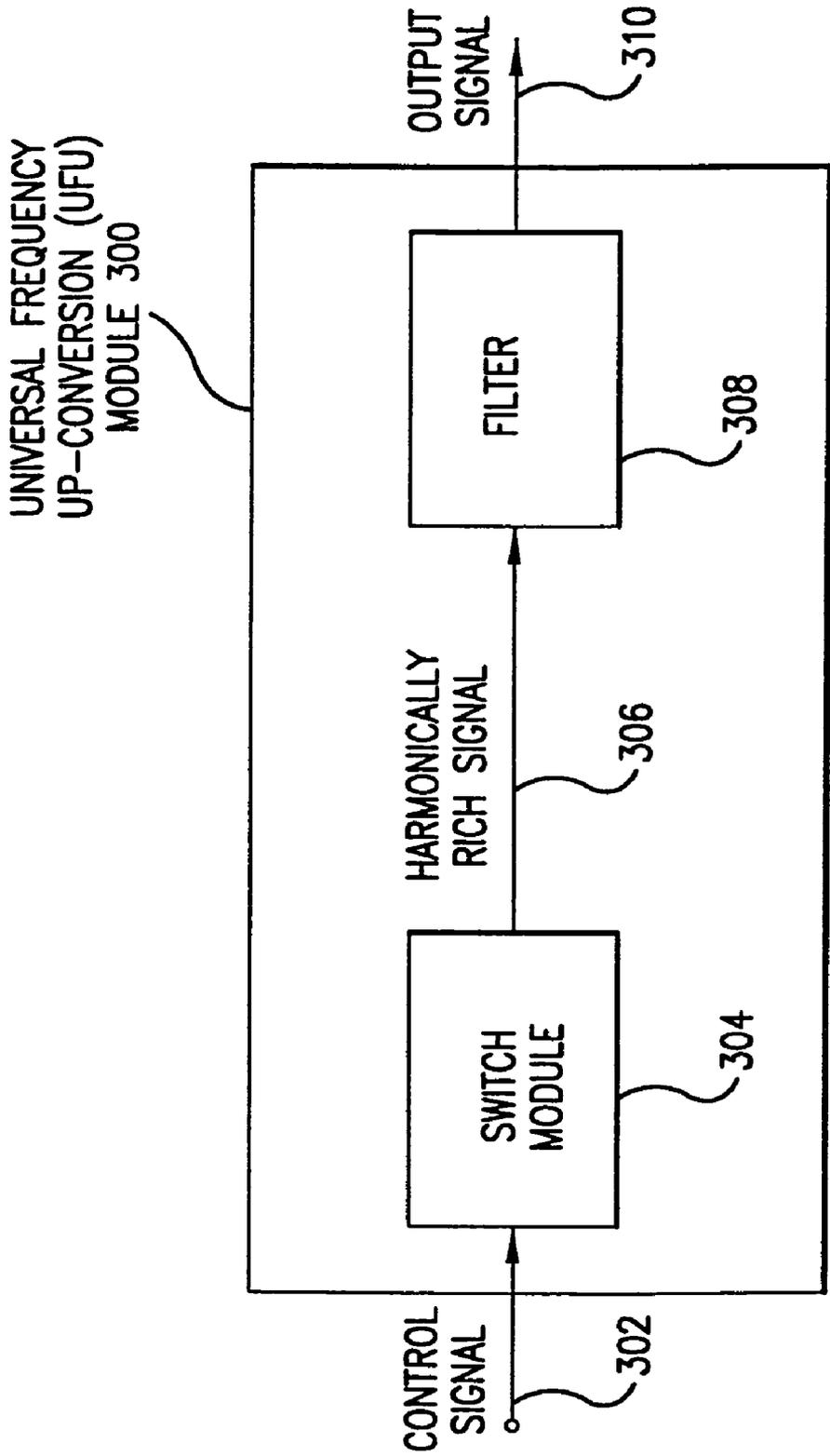


FIG. 3

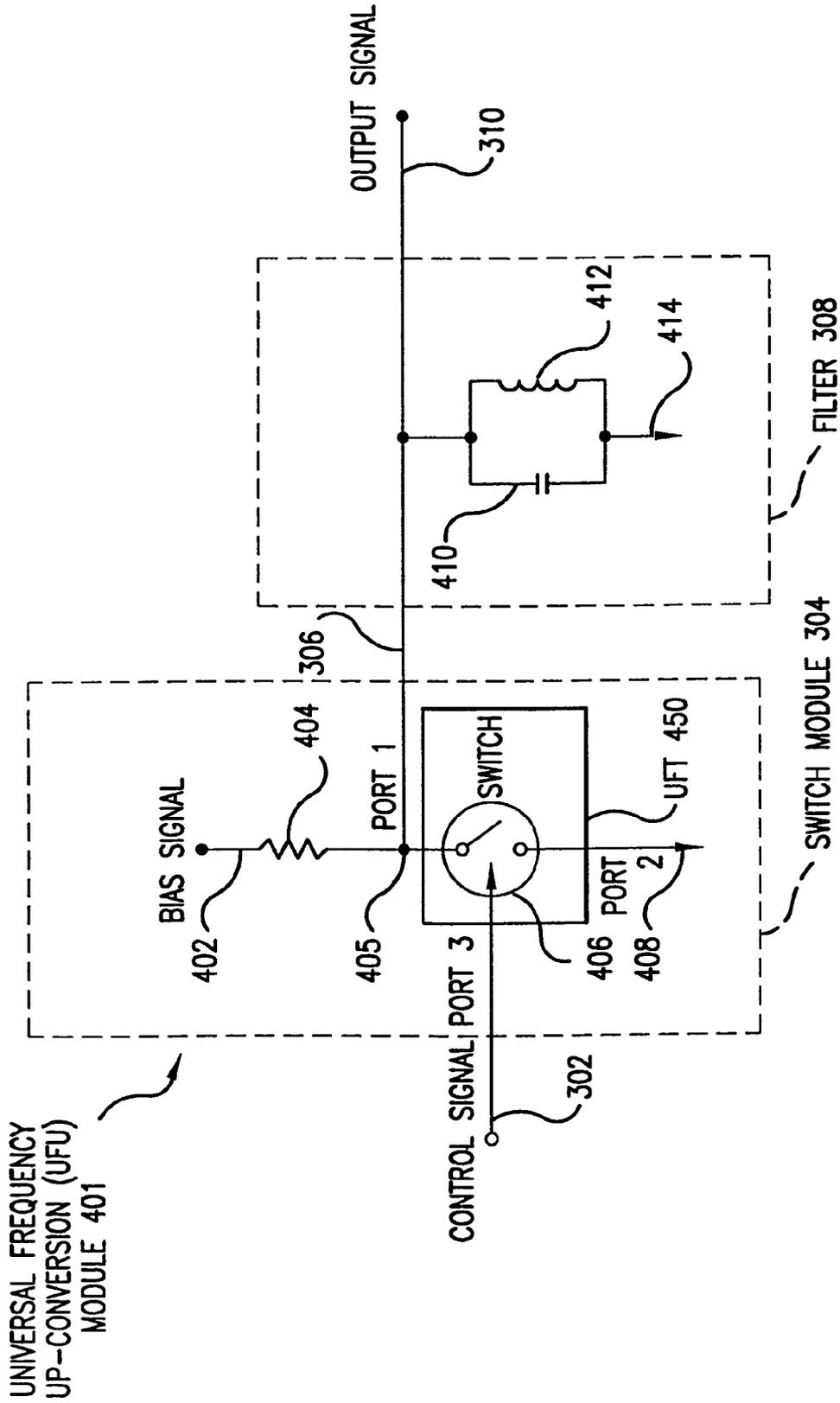


FIG. 4

UNIVERSAL FREQUENCY  
UP-CONVERSION  
(UFU) MODULE 590

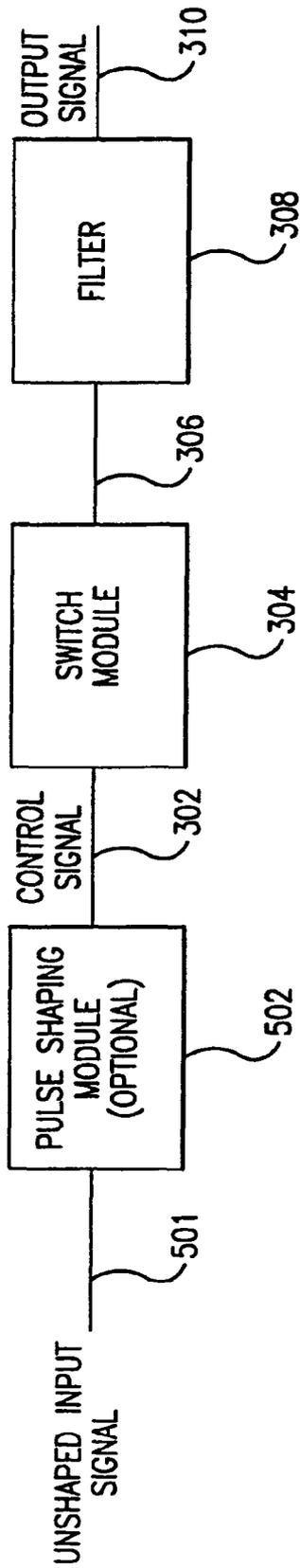
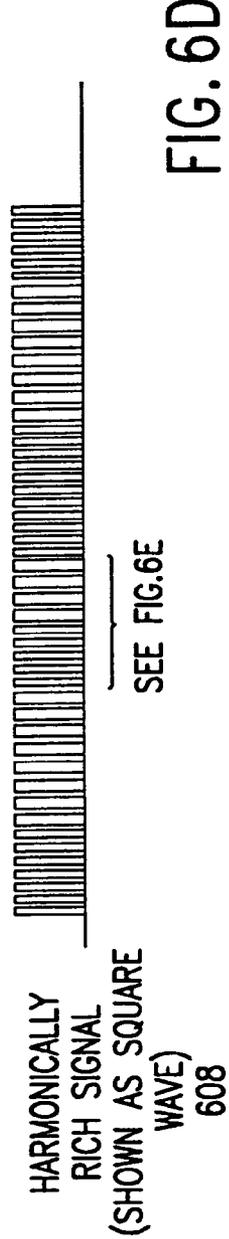
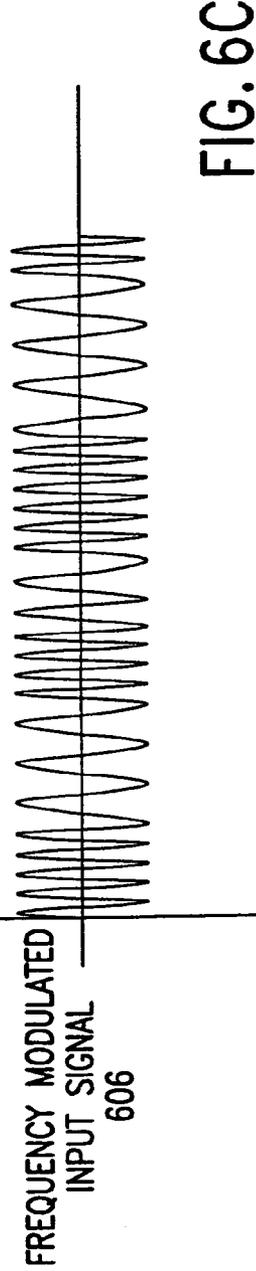
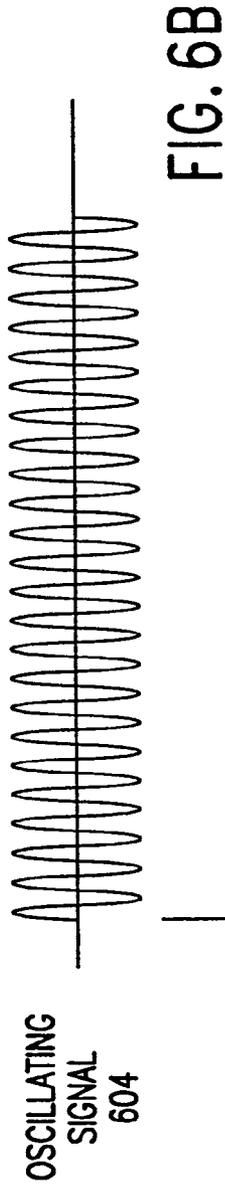
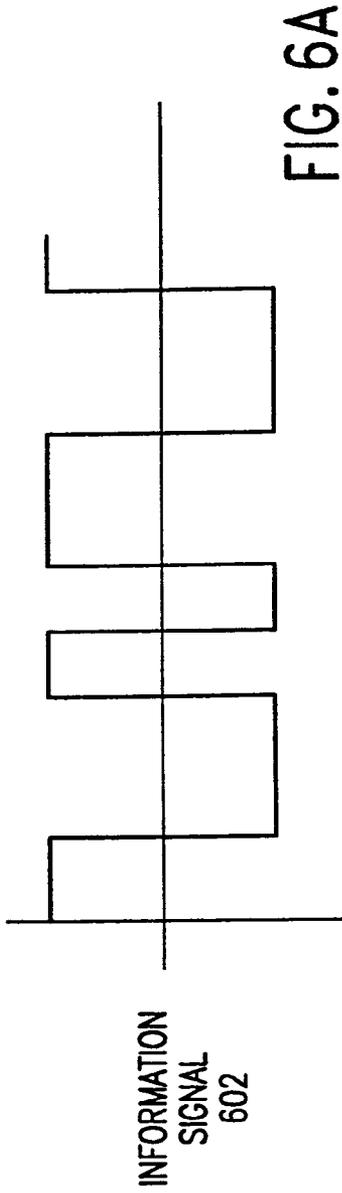
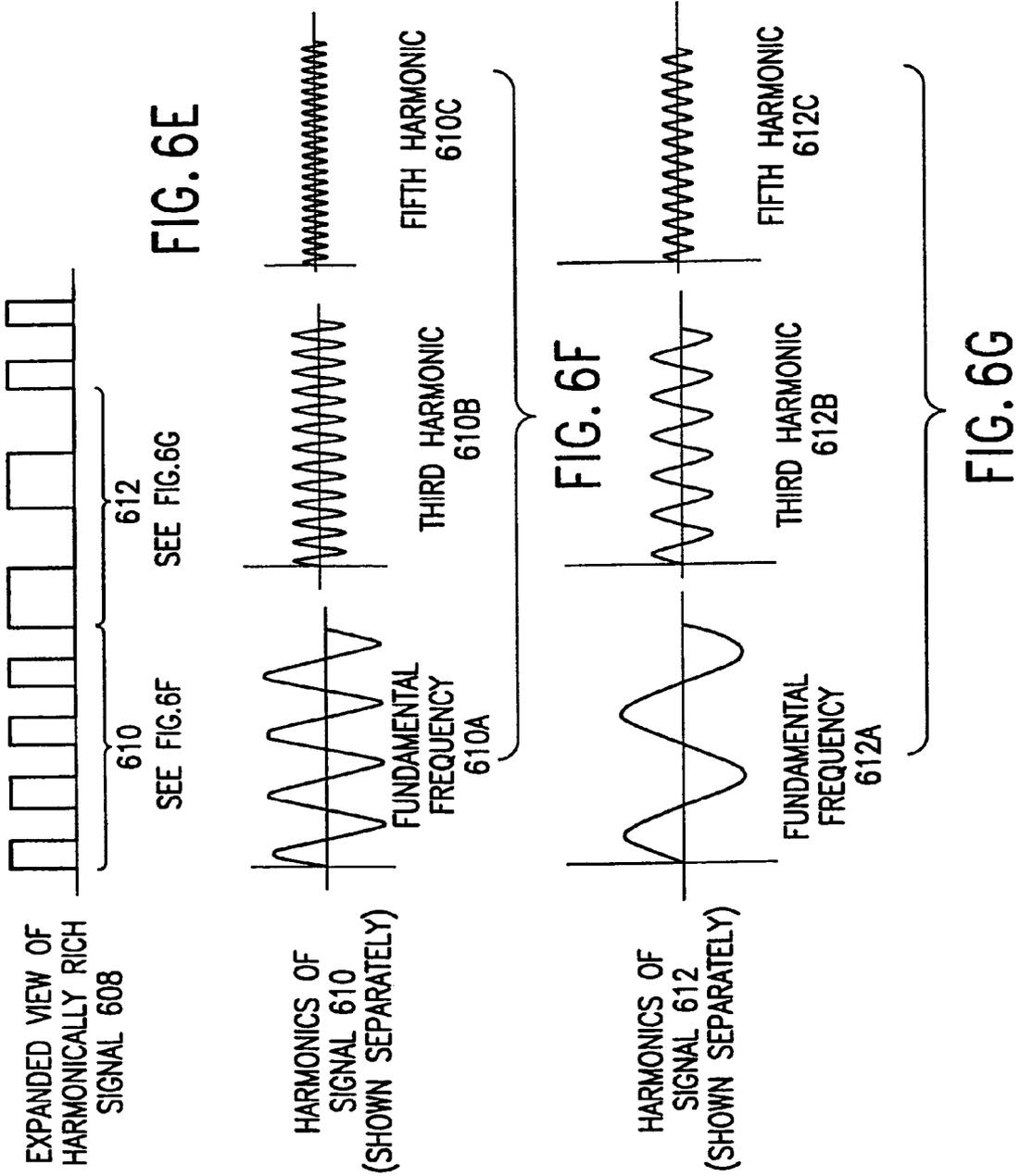
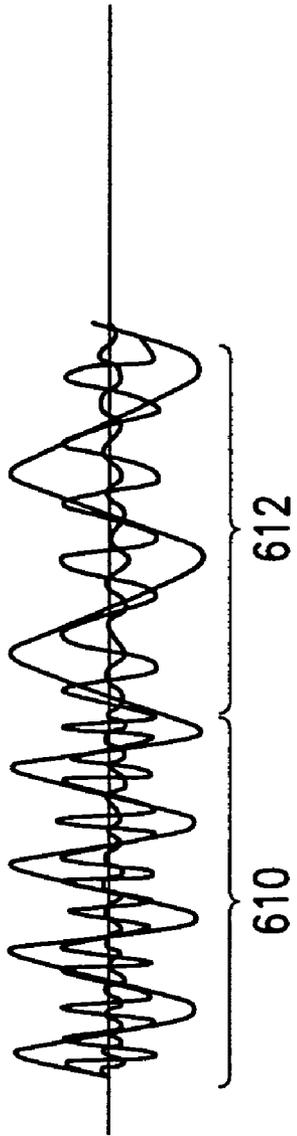


FIG. 5

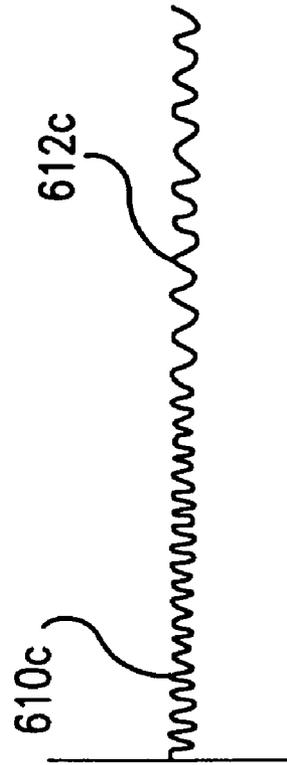






HARMONICS OF  
SIGNALS 610 AND  
612  
(SHOWN SIMULTANEOUSLY  
BUT NOT SUMMED)

FIG. 6H



FILTERED  
OUTPUT  
SIGNAL  
614

FIG. 6I

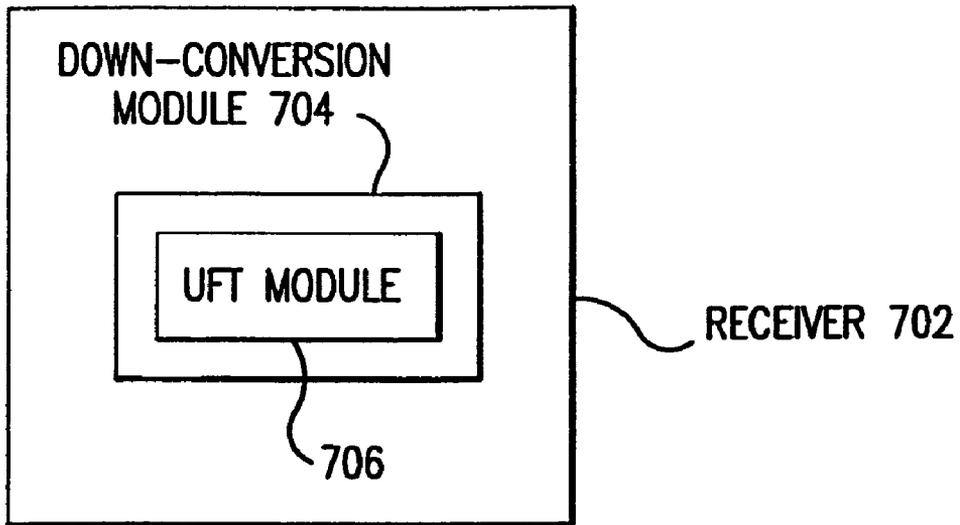


FIG. 7

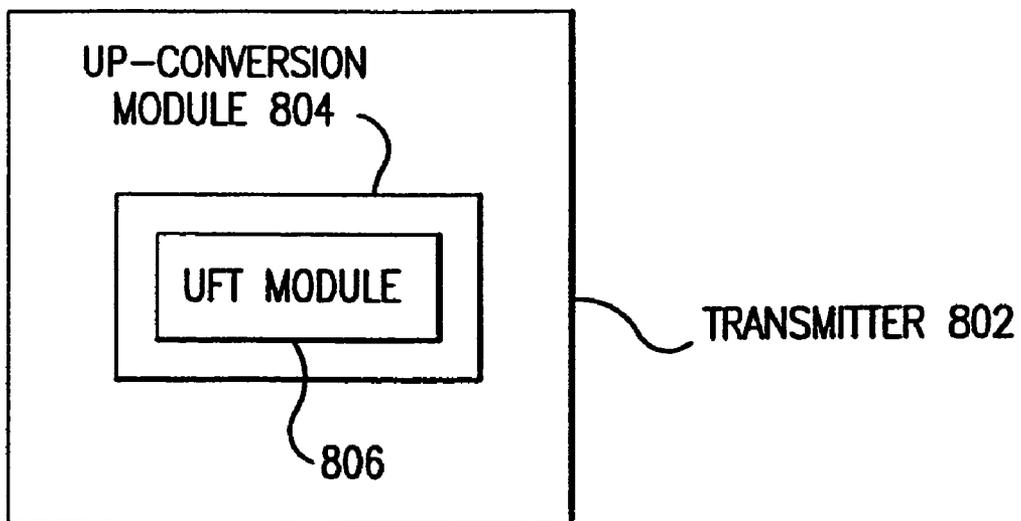


FIG. 8

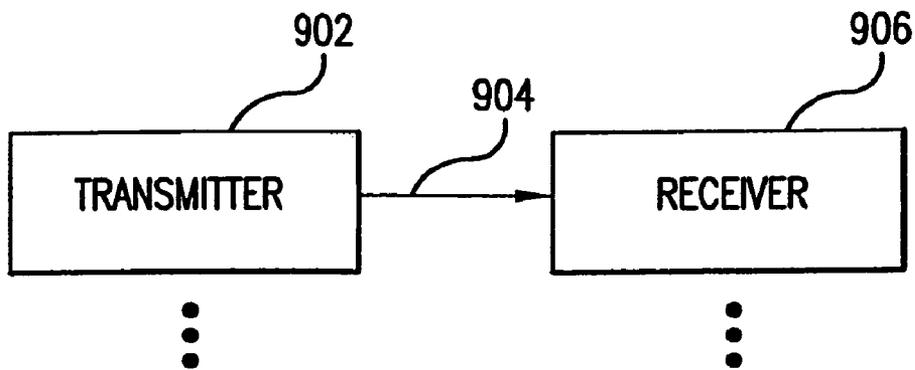


FIG. 9

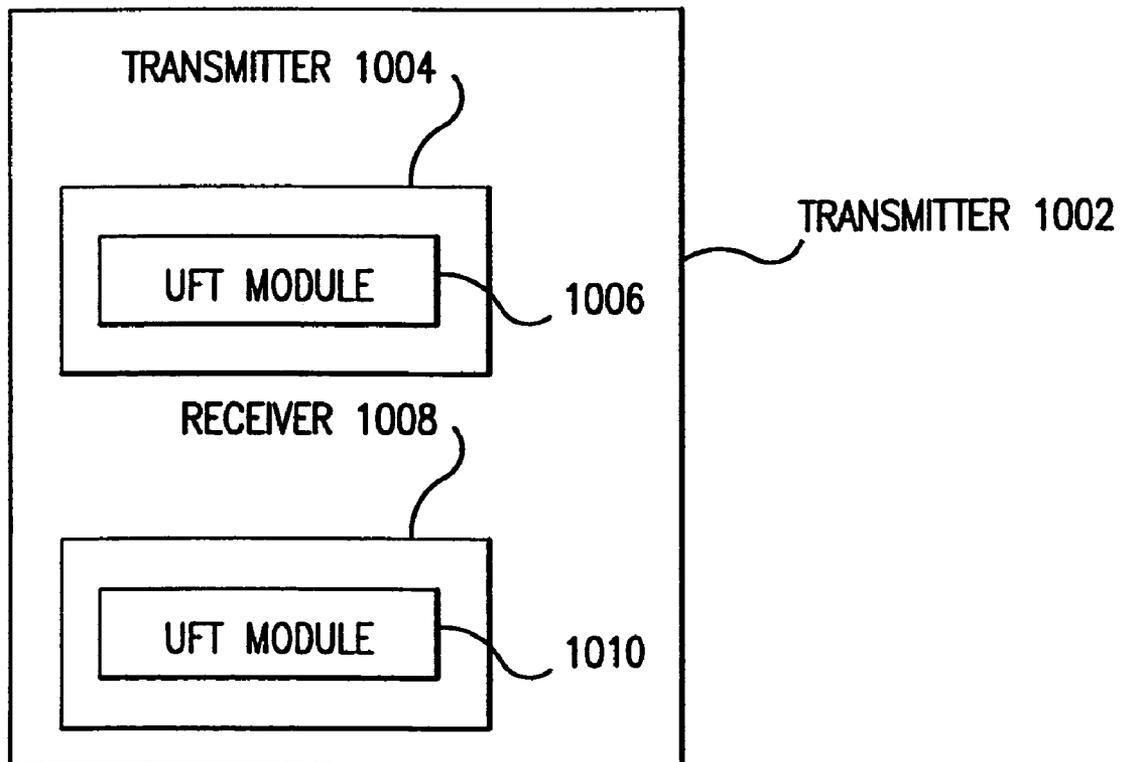


FIG. 10

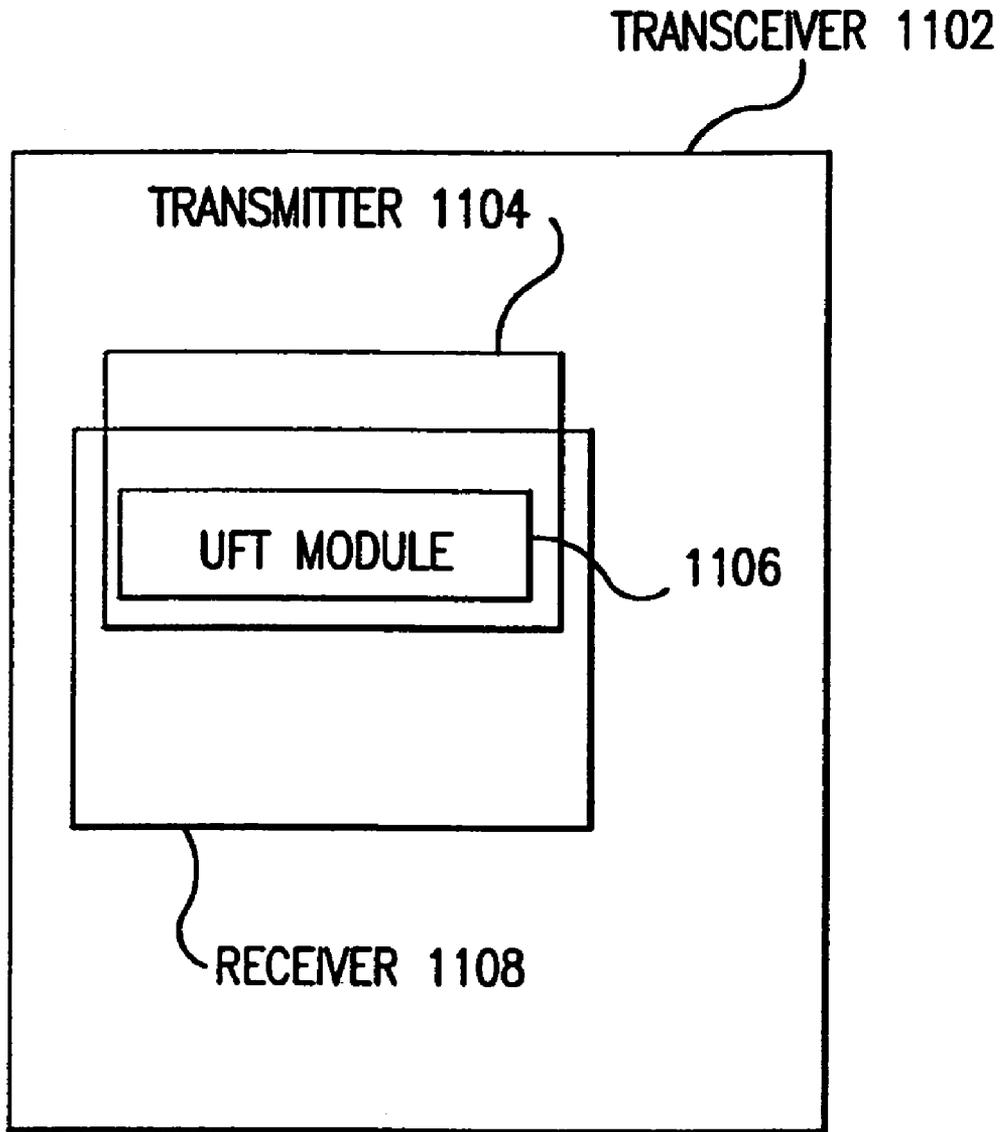


FIG. 11

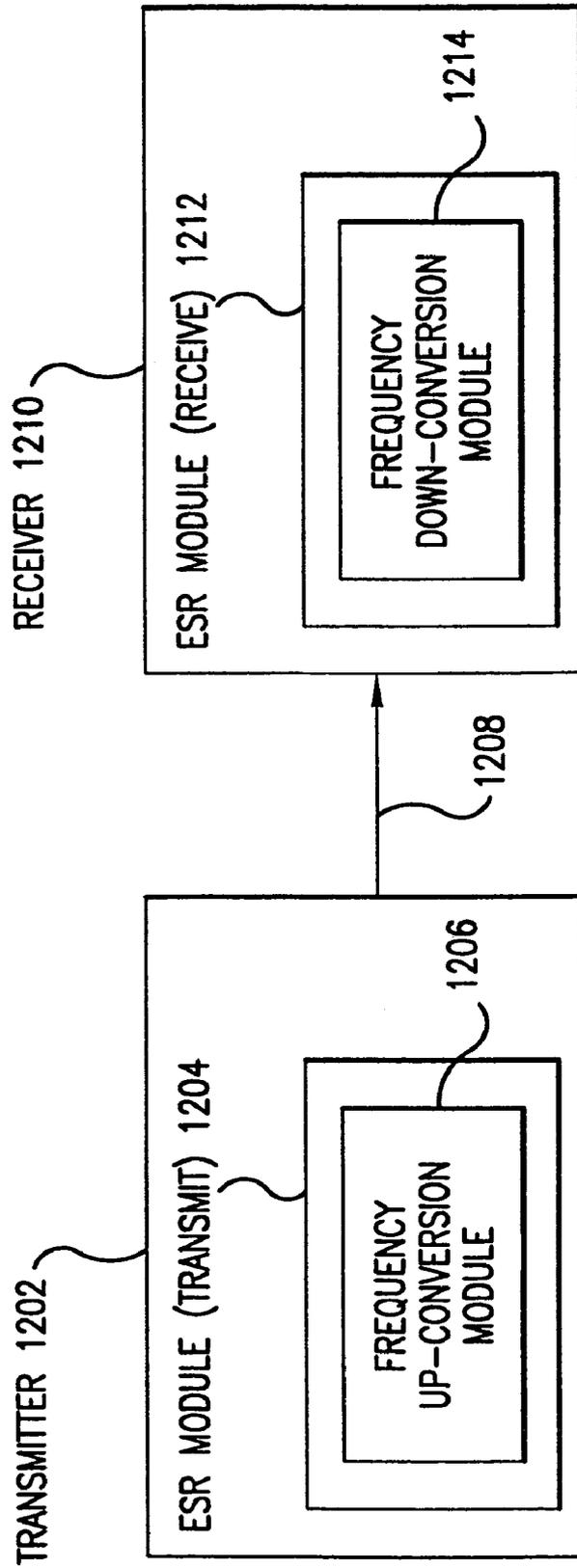


FIG. 12

UNIFIED DOWN-CONVERTING  
AND FILTERING (UDF) MODULE 1302

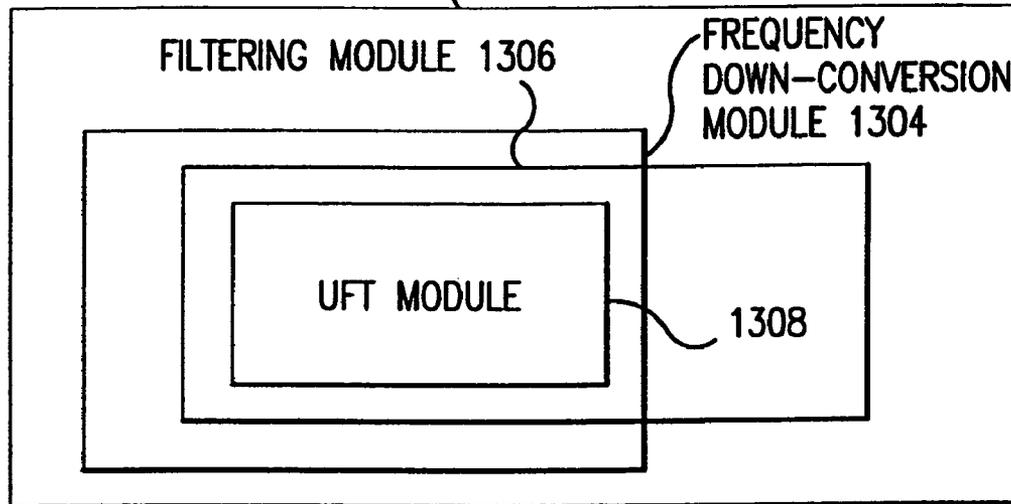


FIG. 13

RECEIVER 1402

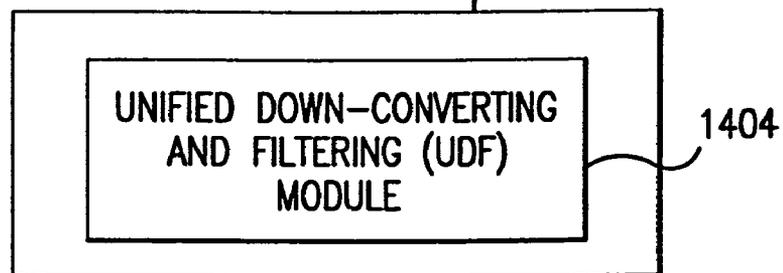


FIG. 14

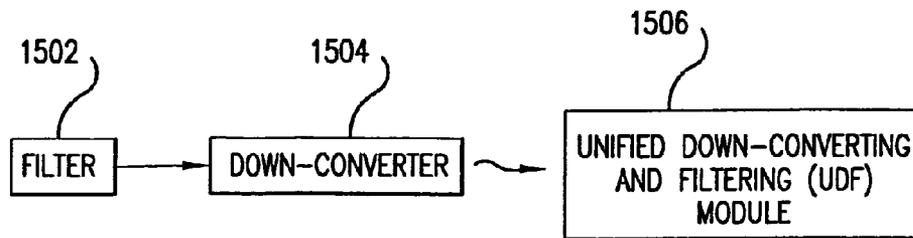


FIG. 15A

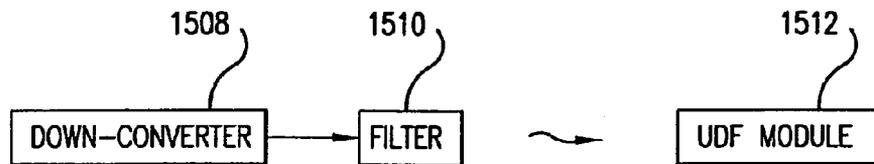


FIG. 15B

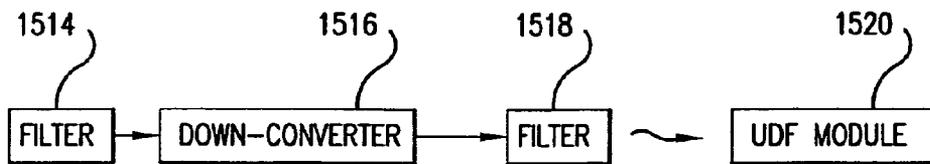


FIG. 15C



FIG. 15D

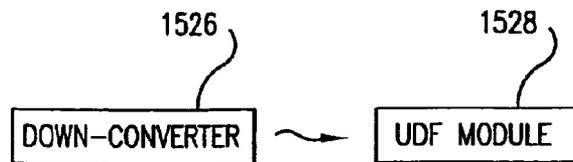


FIG. 15E

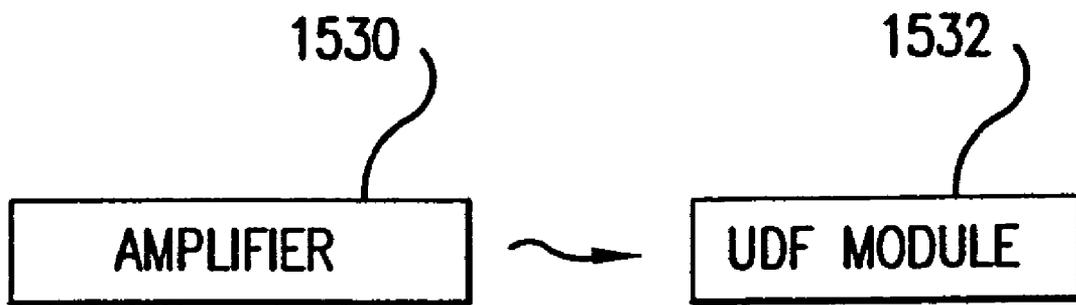


FIG. 15F

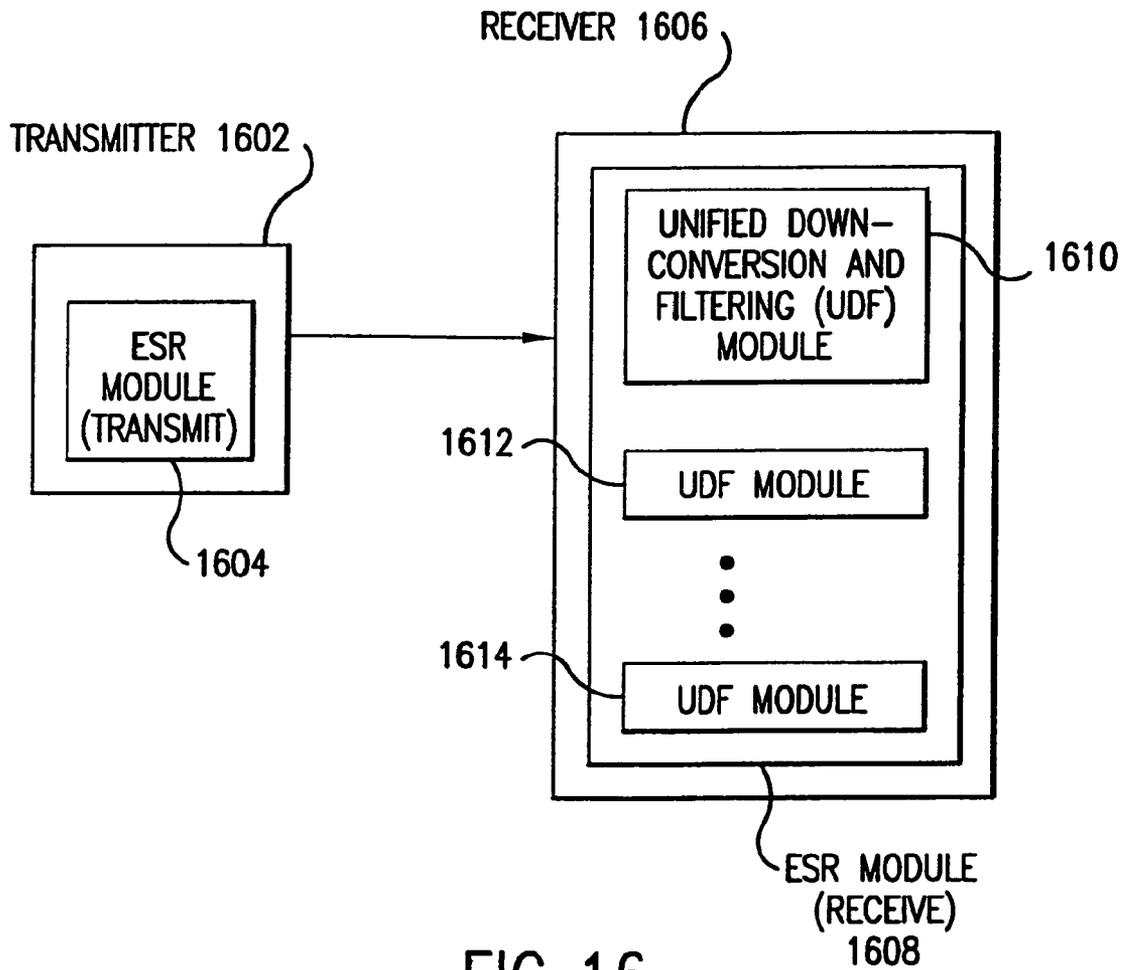


FIG. 16

UNIFIED DOWNCONVERTING AND  
FILTERING (UDF) MODULE 1702

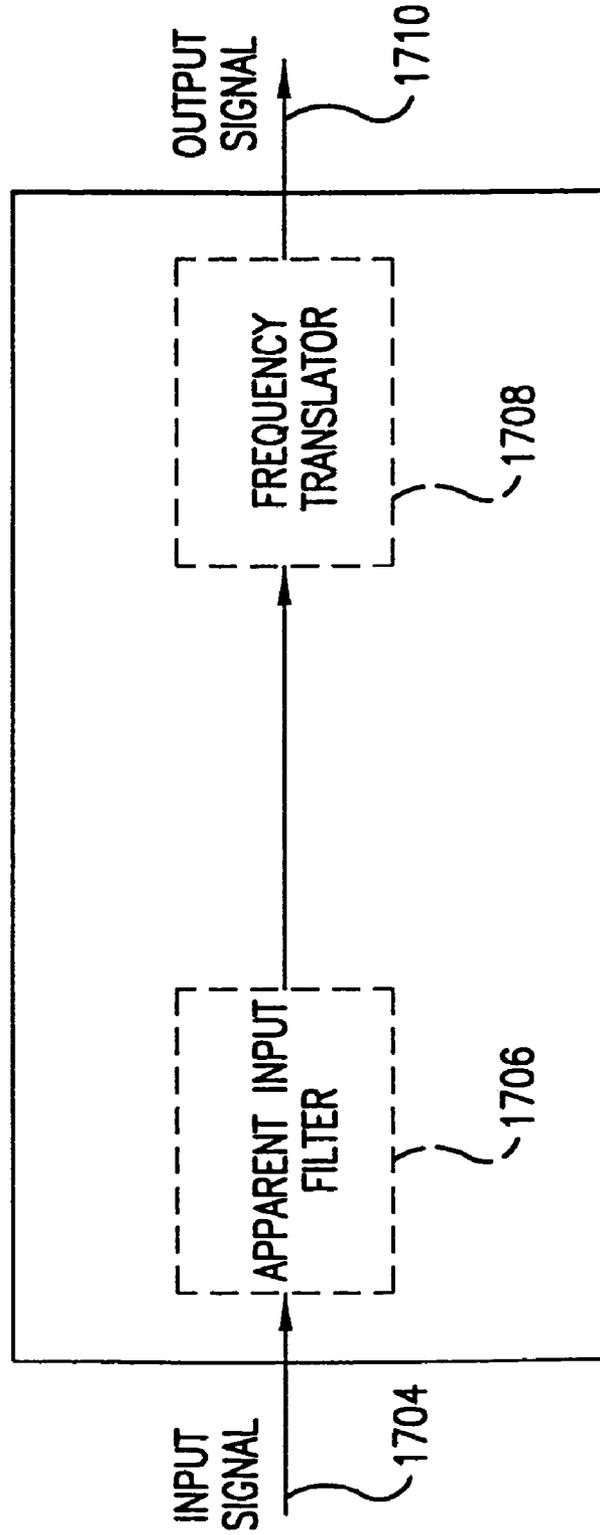


FIG. 17

1802

TIME NODE	t-1 (RISING EDGE OF $\phi_1$ )	t-1 (RISING EDGE OF $\phi_2$ )	t (RISING EDGE OF $\phi_1$ )	t (RISING EDGE OF $\phi_2$ )	t+1 (RISING EDGE OF $\phi_1$ )
1902	$V_{I,t-1}$ 1804	$V_{I,t-1}$ 1808	$V_{I,t}$ 1816	$V_{I,t}$ 1826	$V_{I,t+1}$ 1838
1904	—	$V_{I,t-1}$ 1810	$V_{I,t-1}$ 1818	$V_{I,t}$ 1828	$V_{I,t}$ 1840
1906	$V_{O,t-1}$ 1806	$V_{O,t-1}$ 1812	$V_{O,t}$ 1820	$V_{O,t}$ 1830	$V_{O,t+1}$ 1842
1908	—	$V_{O,t-1}$ 1814	$V_{O,t-1}$ 1822	$V_{O,t}$ 1832	$V_{O,t}$ 1844
1910	— 1807	—	$V_{O,t-1}$ 1824	$V_{O,t-1}$ 1834	$V_{O,t}$ 1846
1912	—	— 1815	—	$V_{O,t-1}$ 1836	$V_{O,t-1}$ 1848
1918	—	—	—	—	$V_{I,t-}$ 1850 $0.1*V_{O,t-}$ $0.8*V_{O,t-1}$

FIG. 18

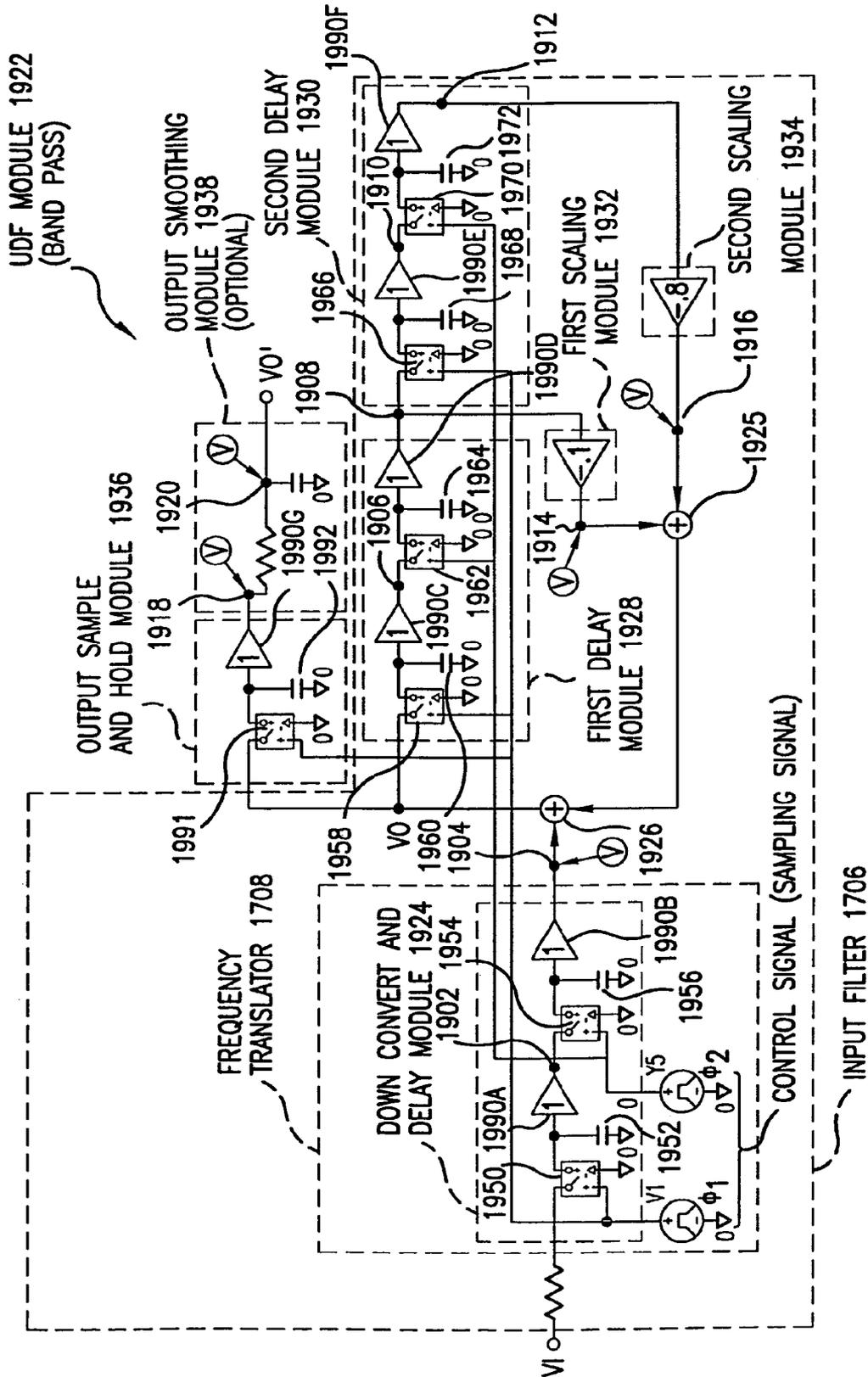


FIG. 19

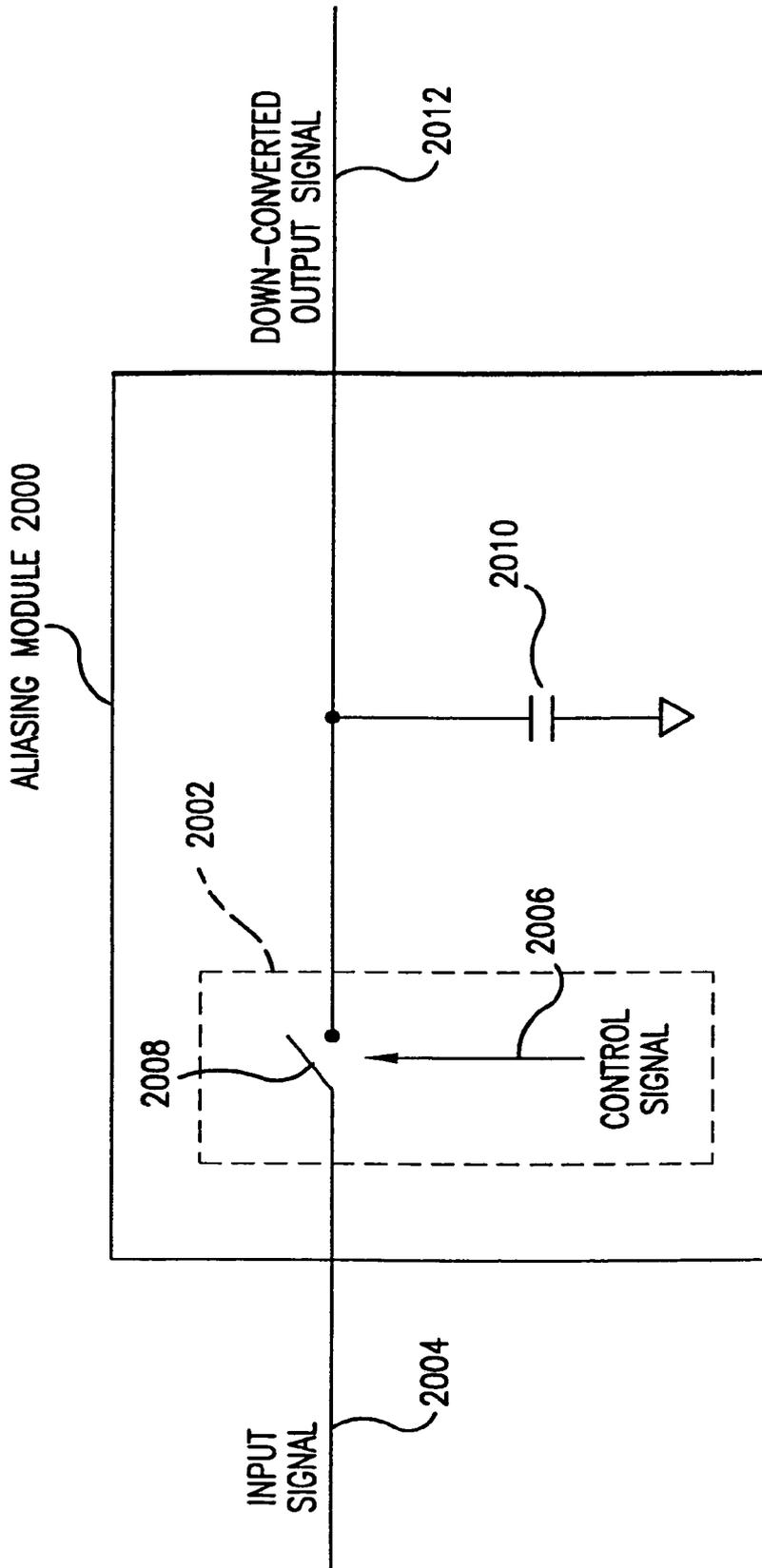


FIG. 20A

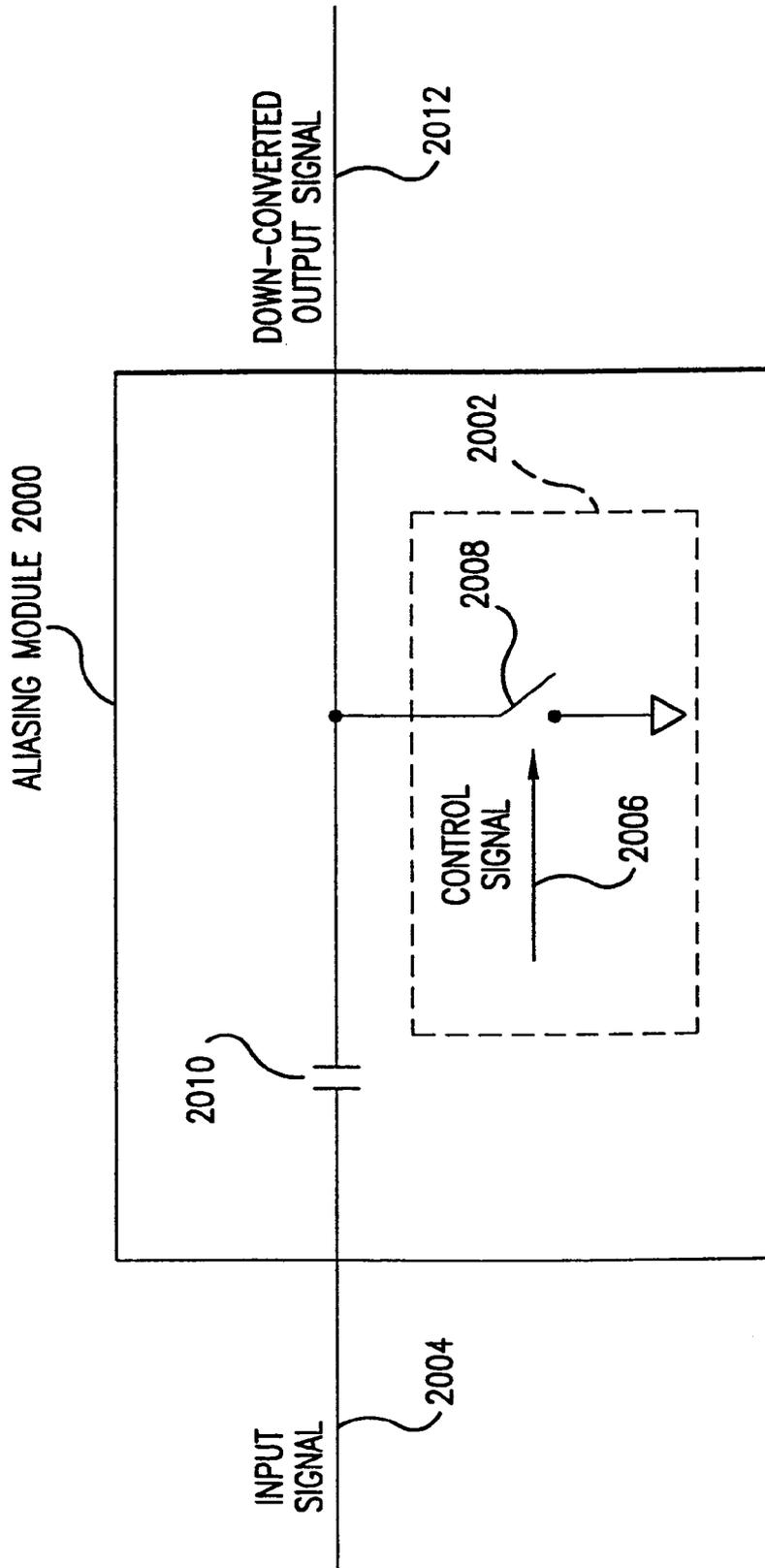


FIG. 20A-1

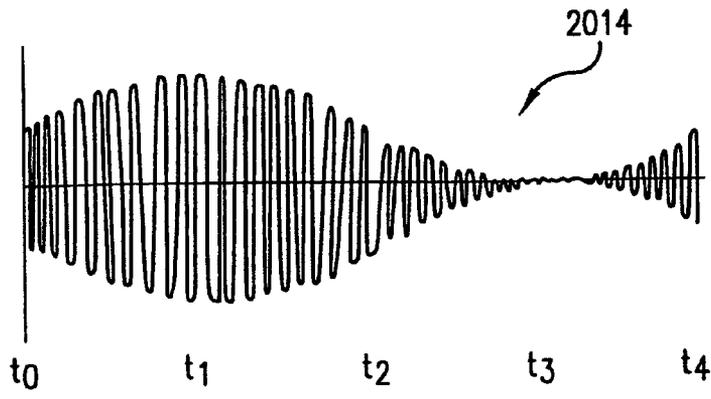


FIG. 20B

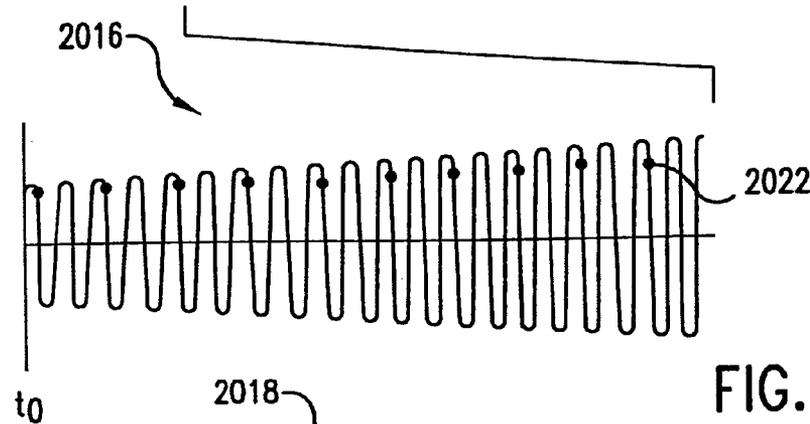


FIG. 20C



FIG. 20D

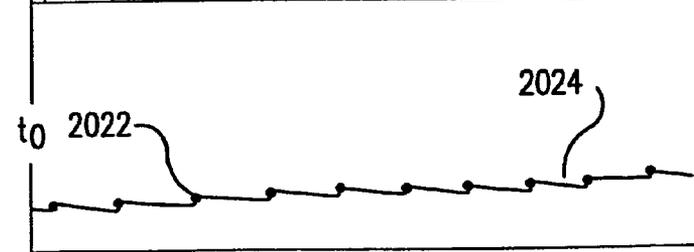


FIG. 20E

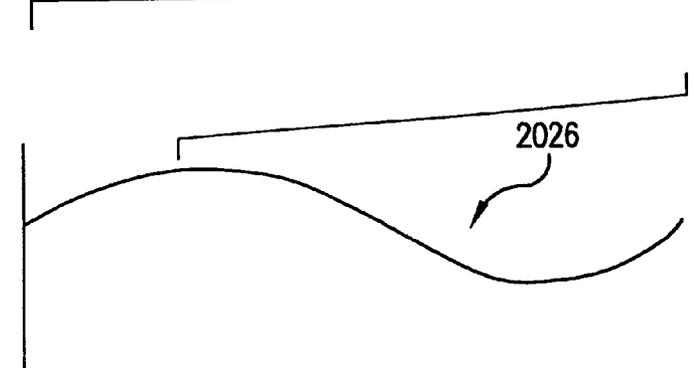


FIG. 20F

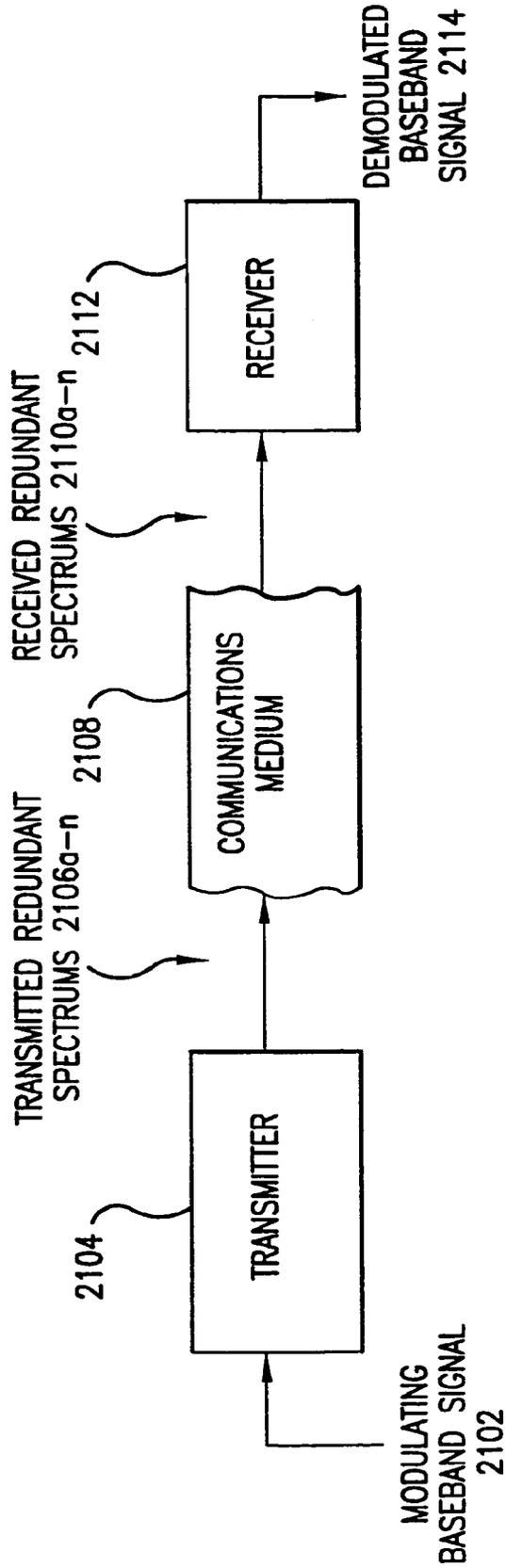


FIG. 21



FIG. 22A

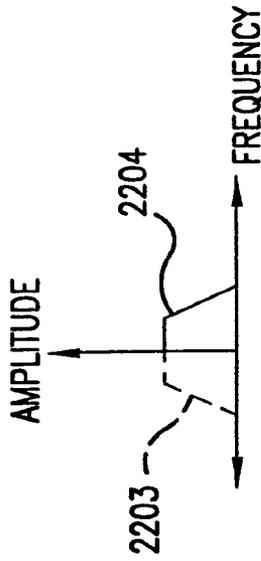


FIG. 22B

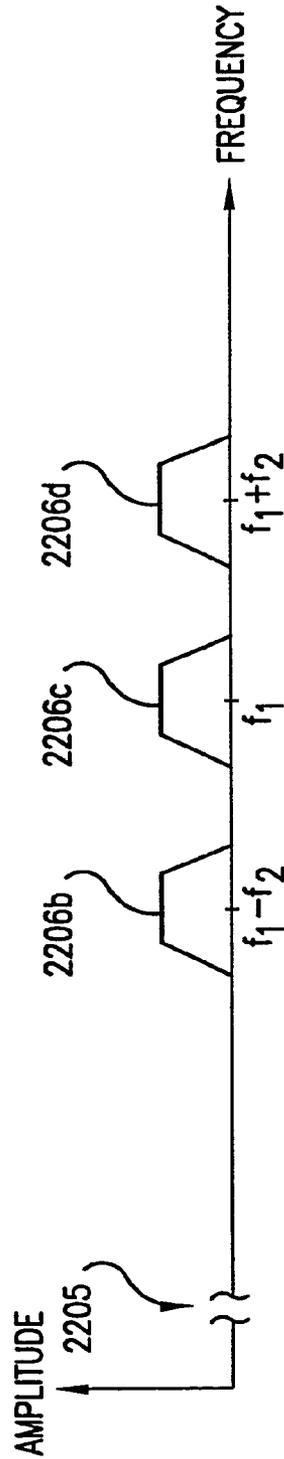


FIG. 22C

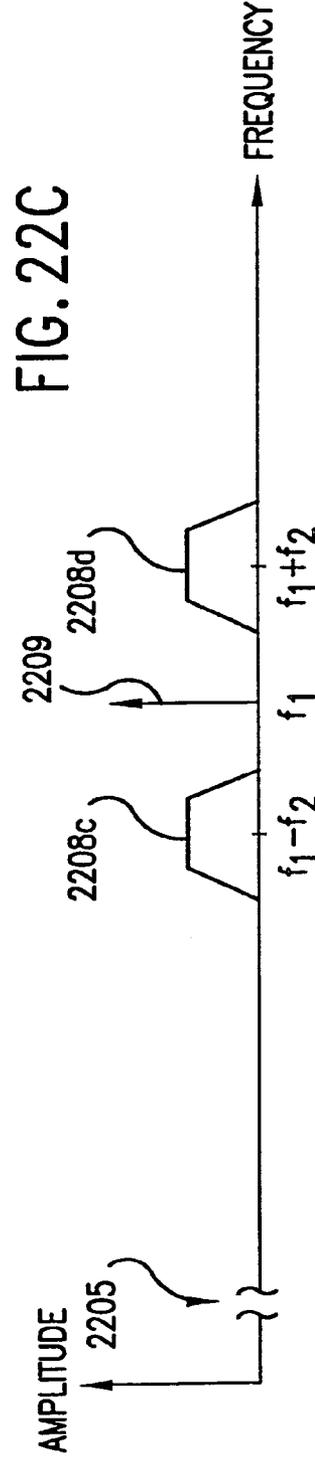


FIG. 22D

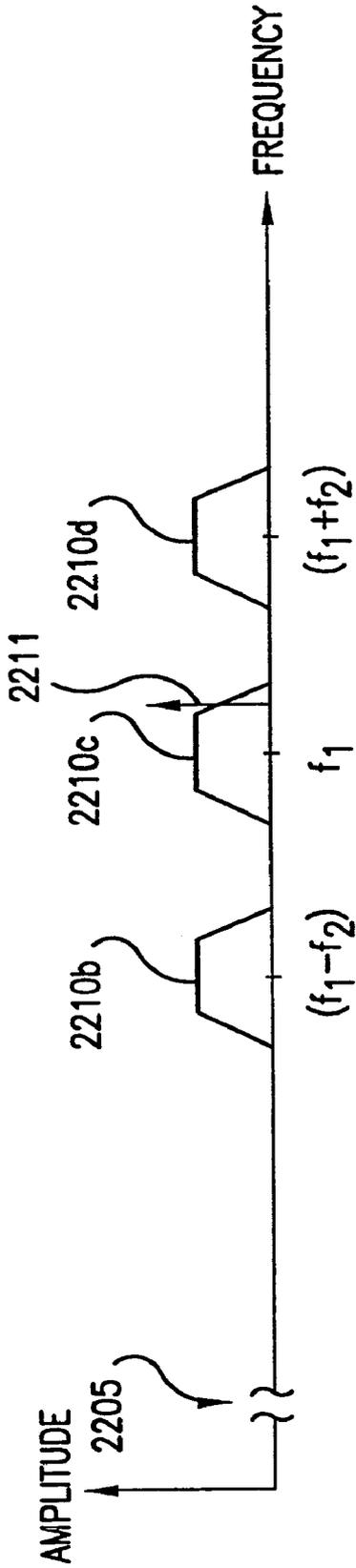


FIG. 22E

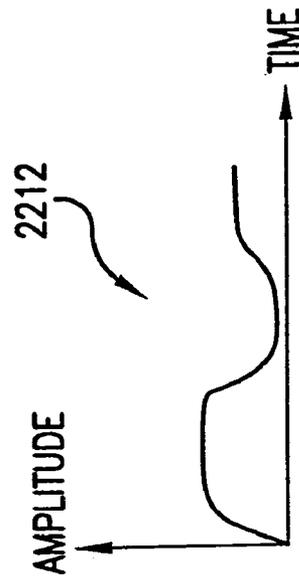


FIG. 22F

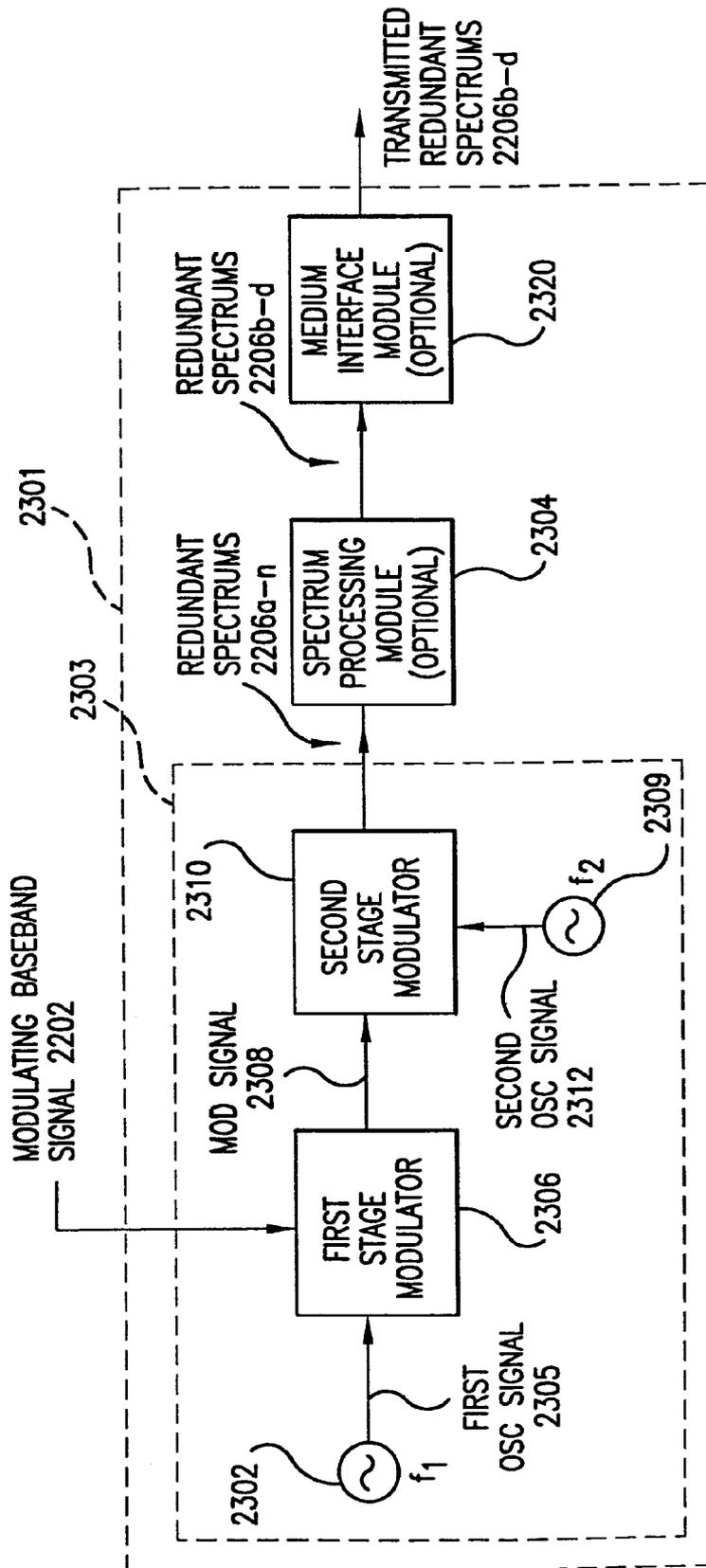


FIG. 23A

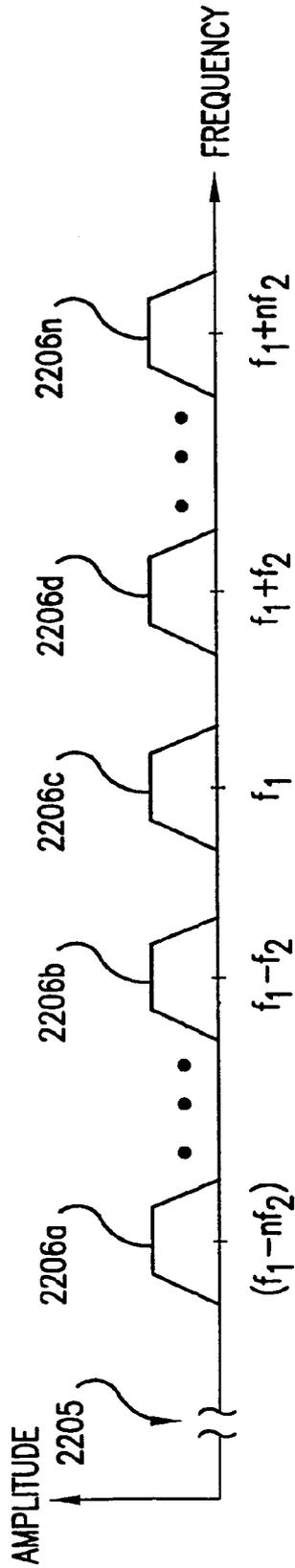


FIG. 23B

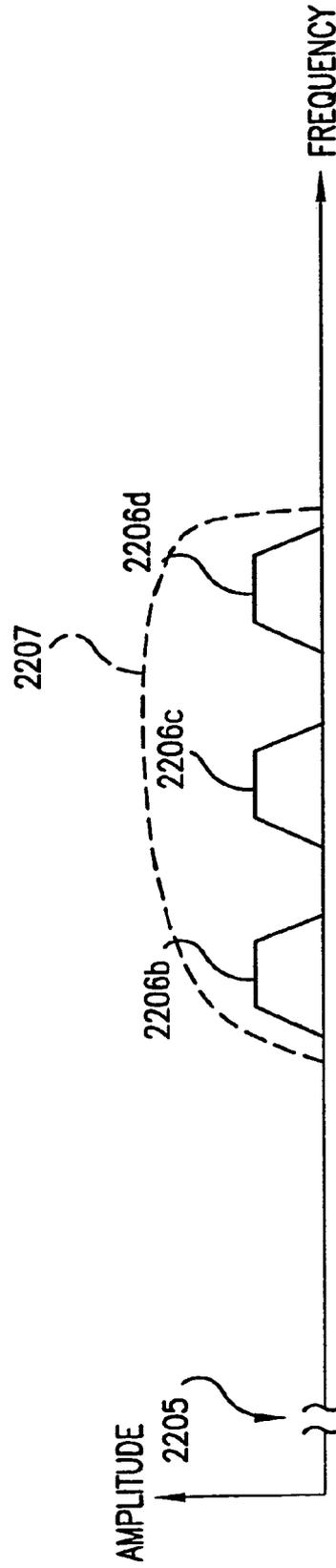


FIG. 23C

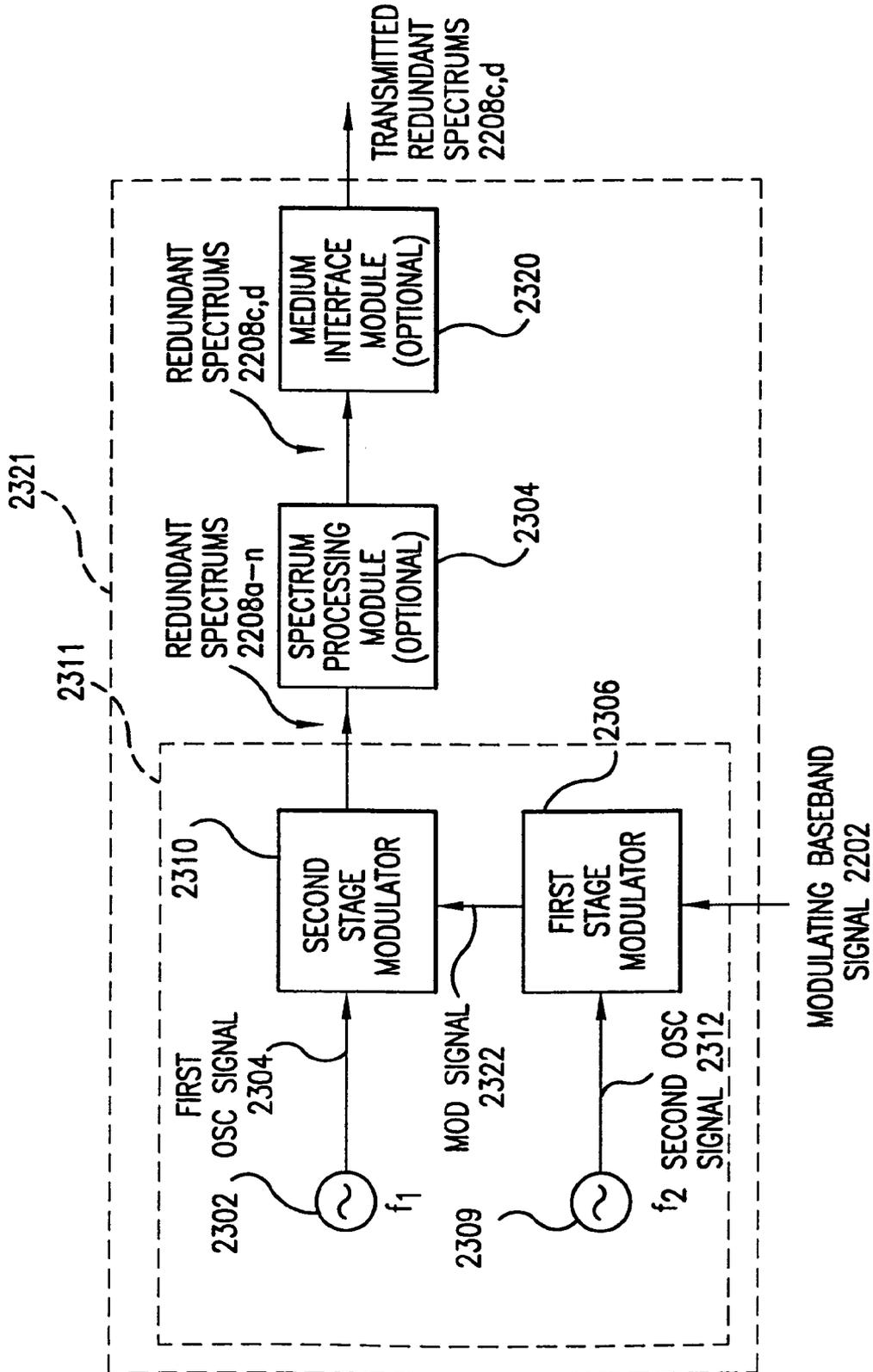


FIG. 230D

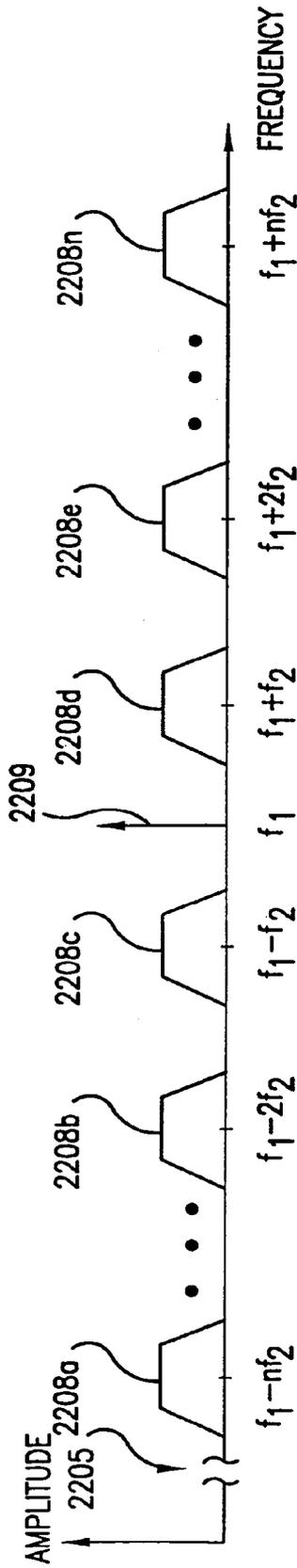


FIG. 23E

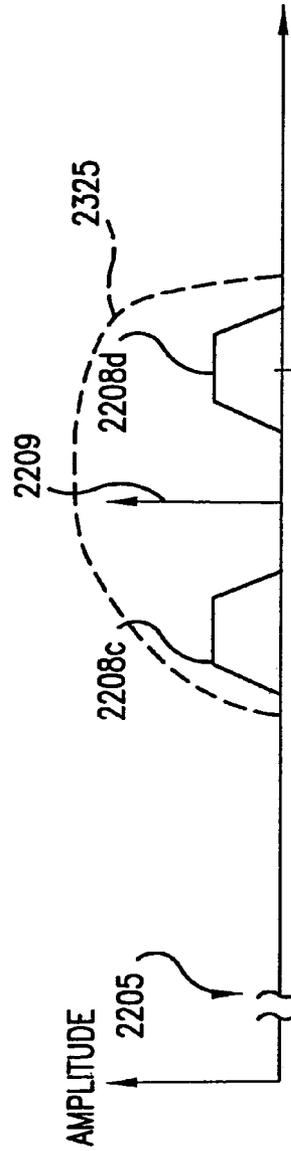


FIG. 23F

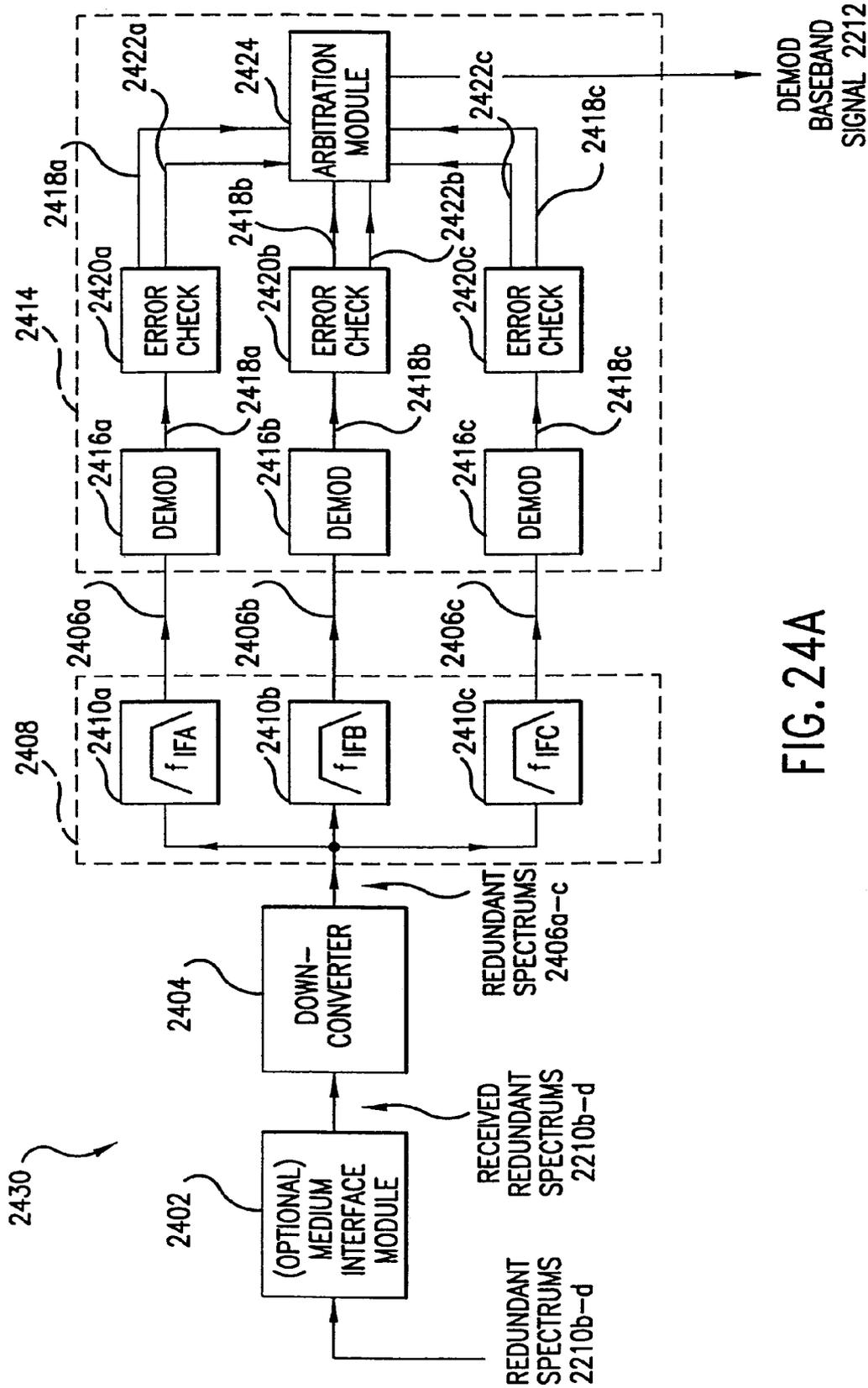


FIG. 24A

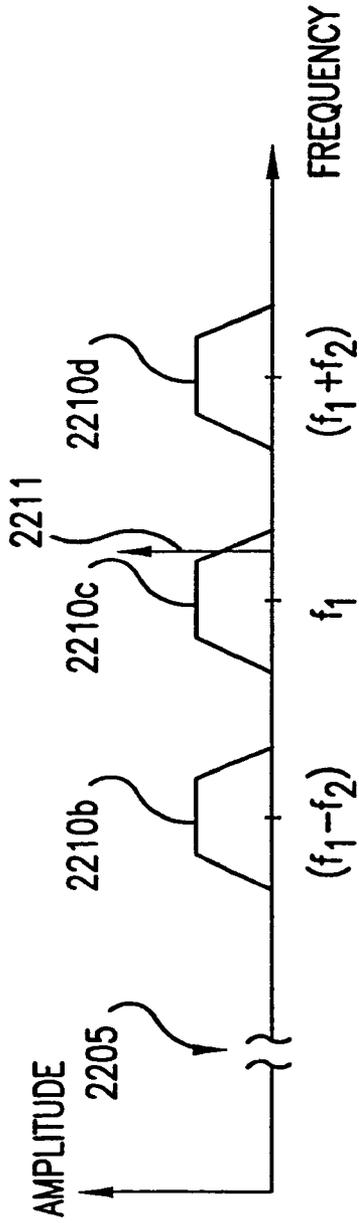


FIG. 24B

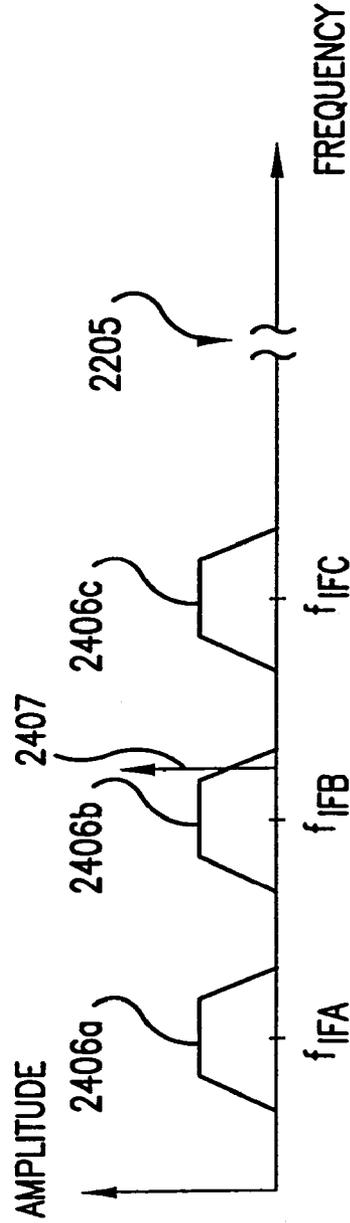


FIG. 24C

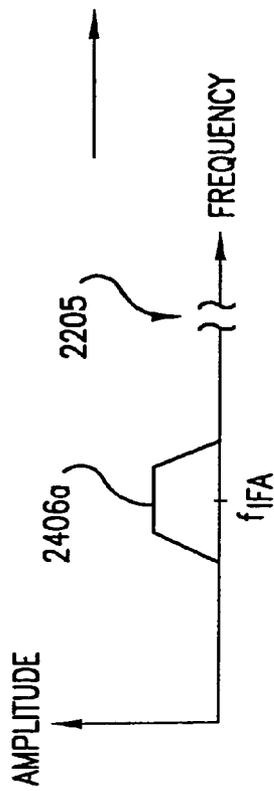


FIG. 24D

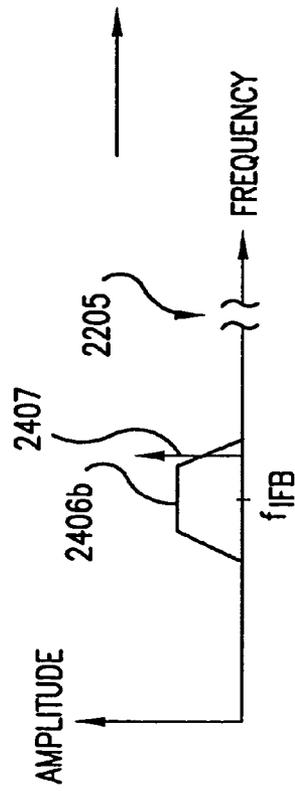


FIG. 24E

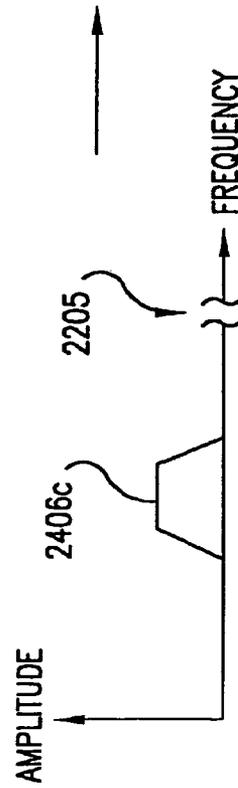


FIG. 24F

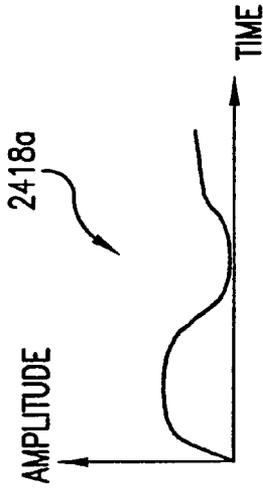


FIG. 24G

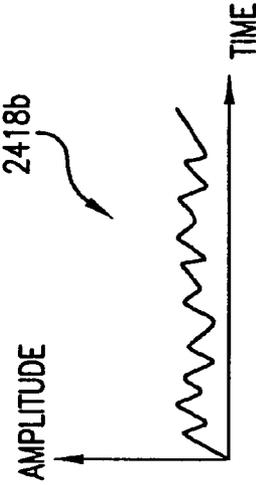


FIG. 24H

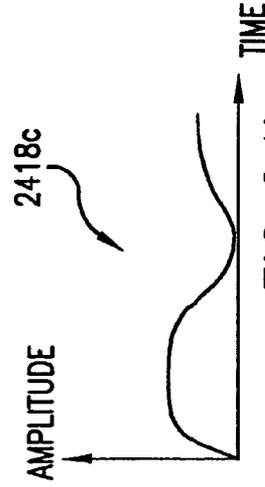


FIG. 24I

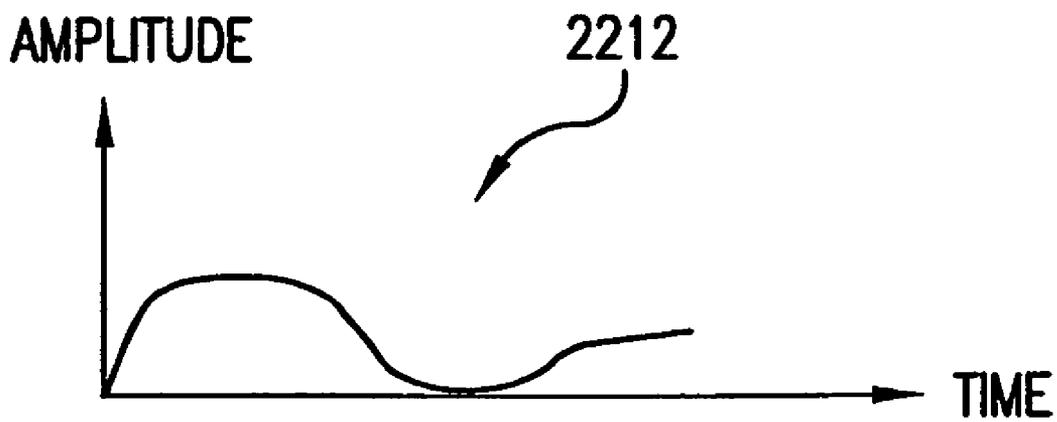


FIG. 24J

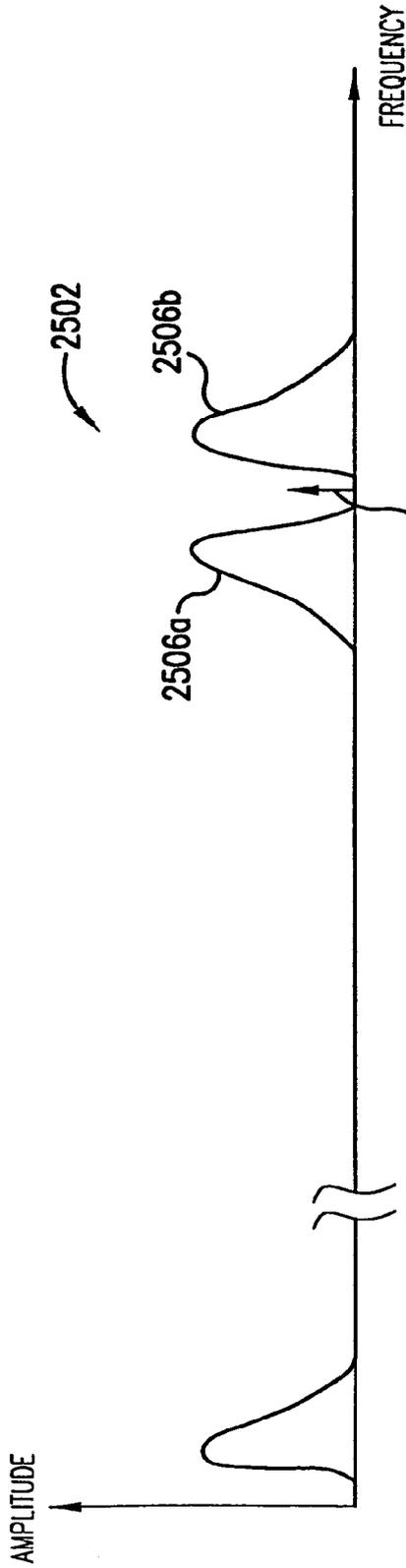


FIG. 25A

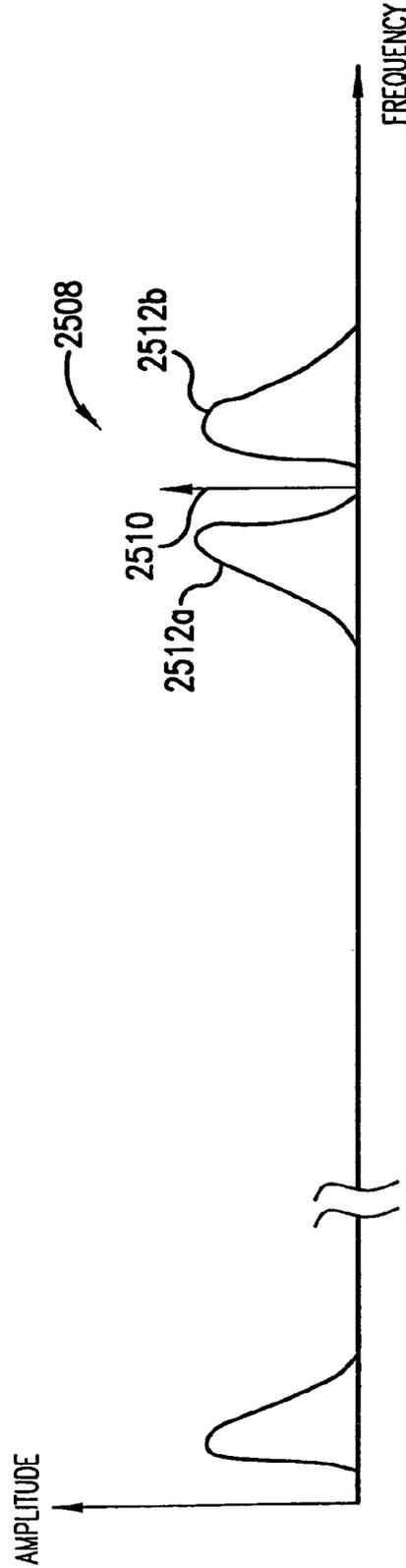


FIG. 25B

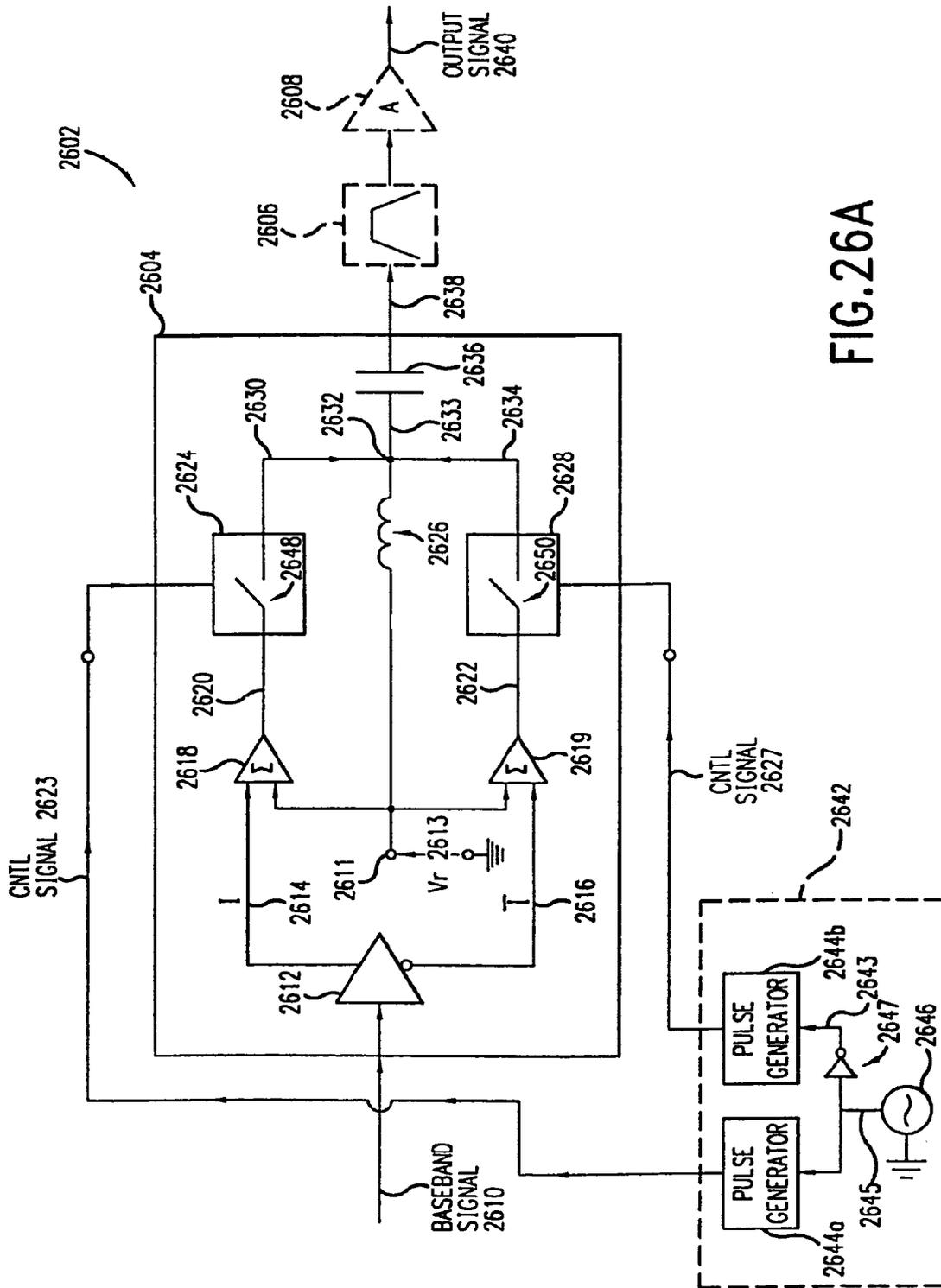


FIG. 26A

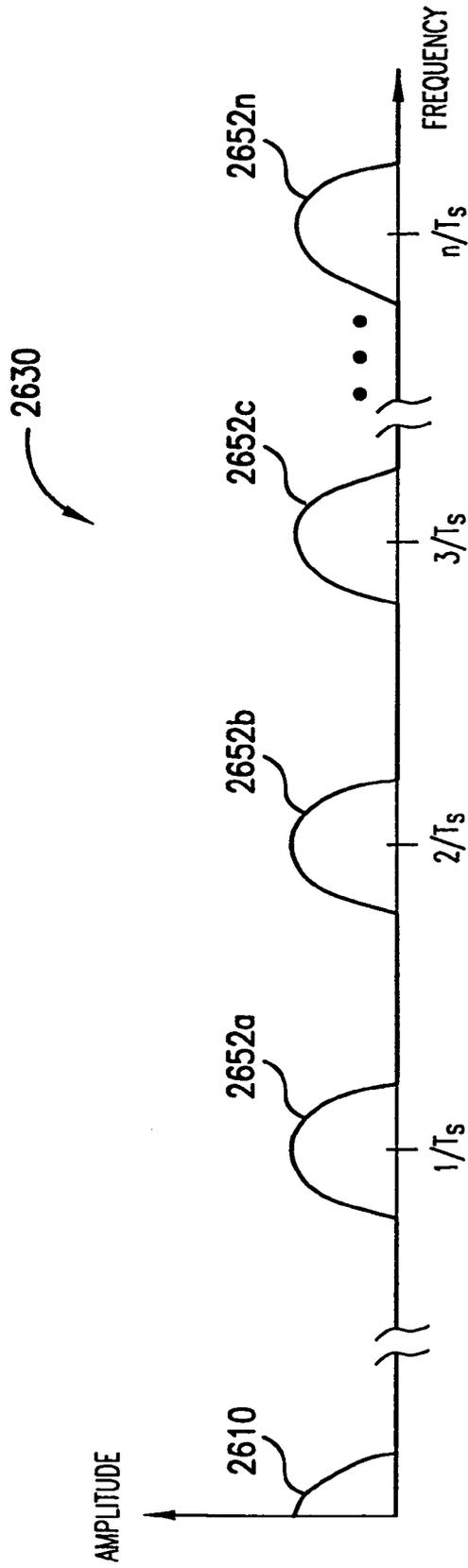


FIG. 26B

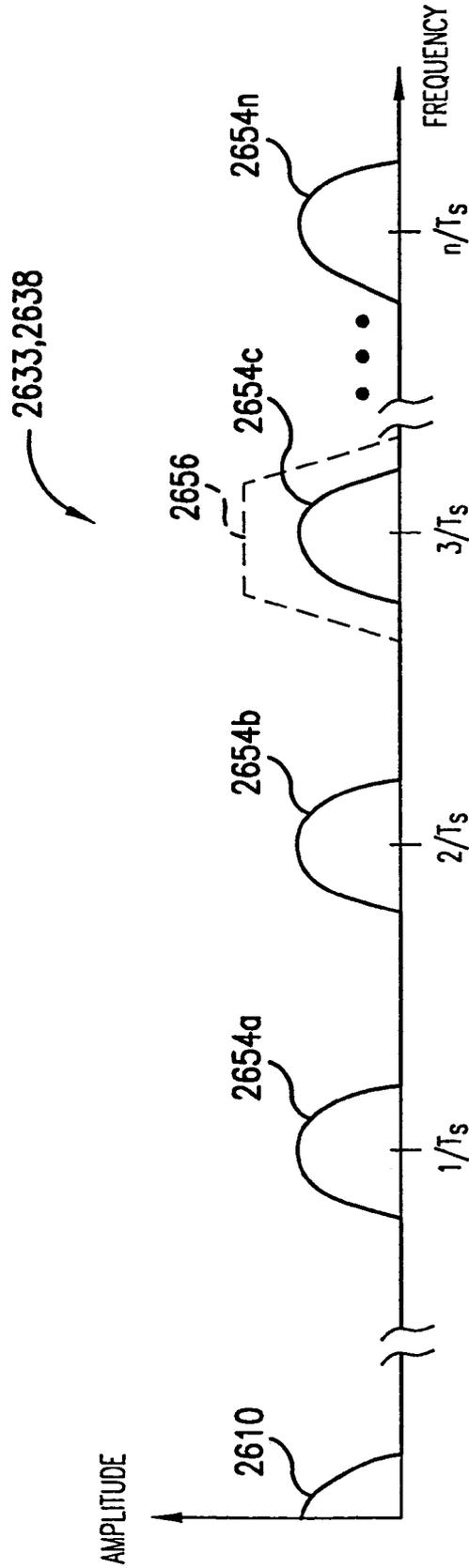


FIG. 26C

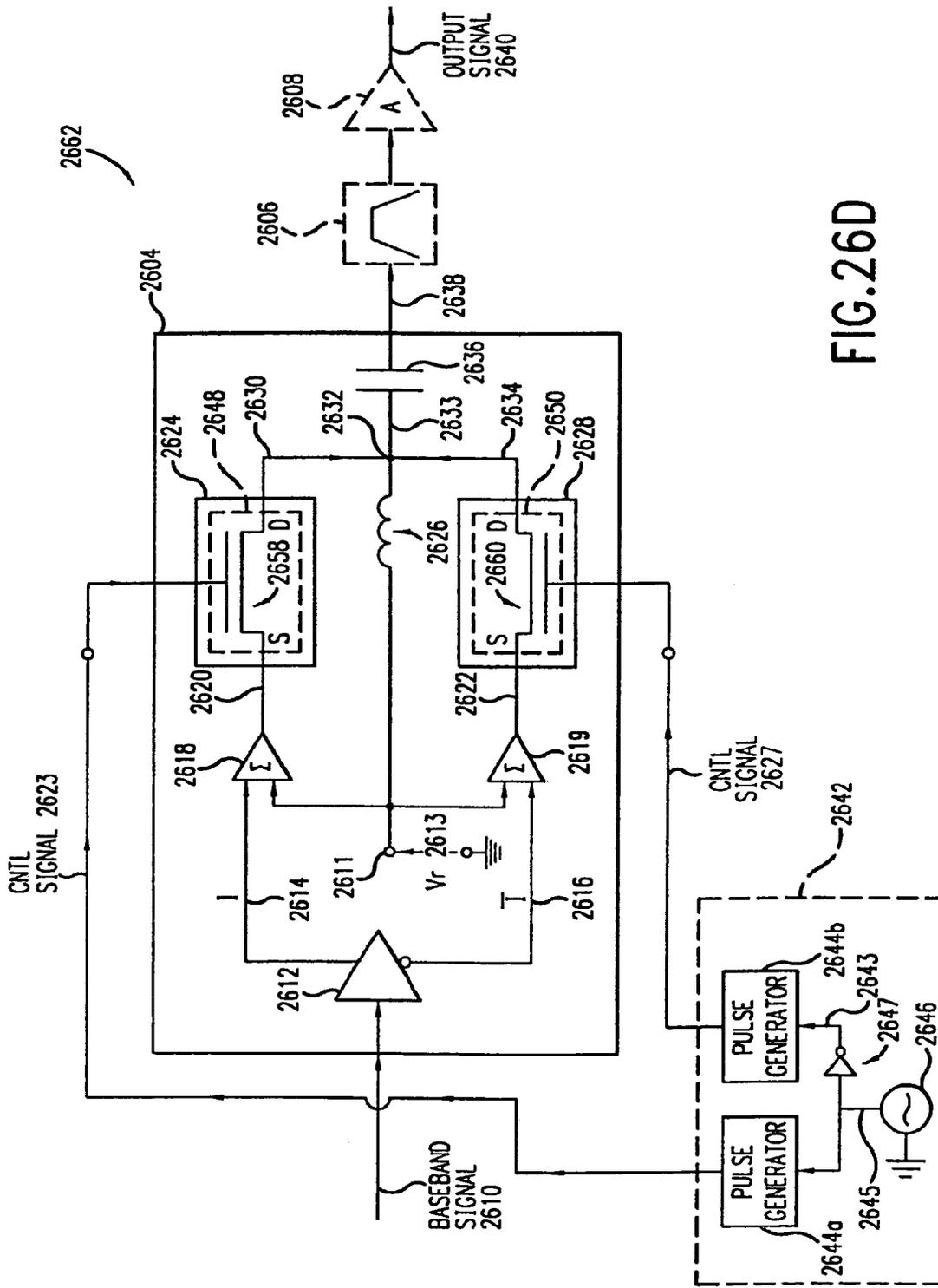


FIG. 26D

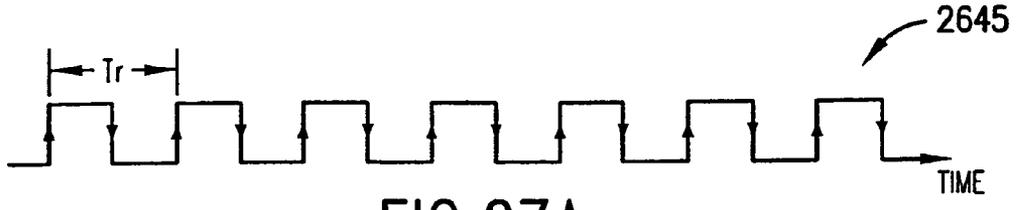


FIG. 27A

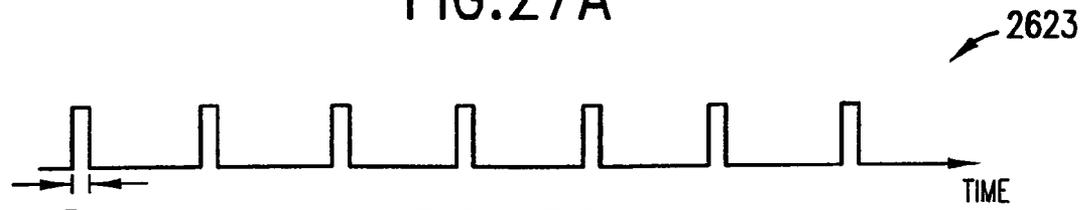


FIG. 27B



FIG. 27C

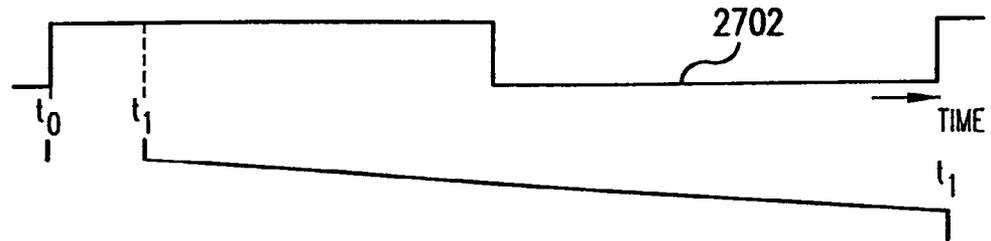


FIG. 27D

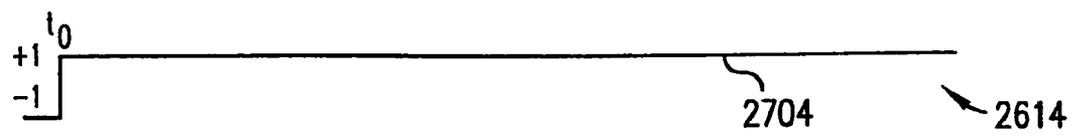


FIG. 27E

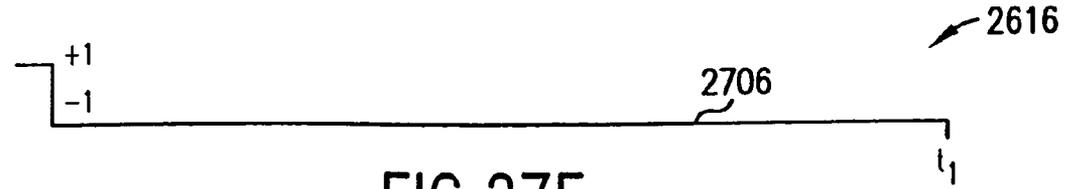


FIG. 27F

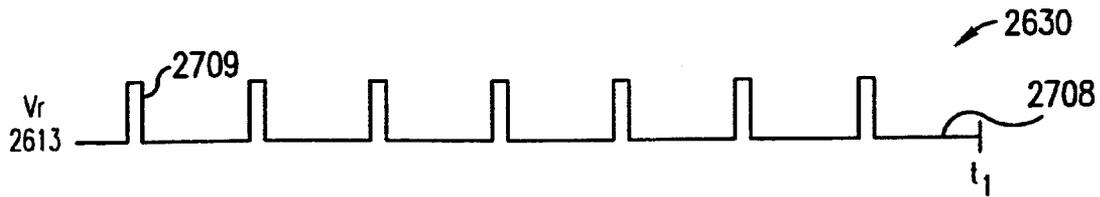


FIG. 27G

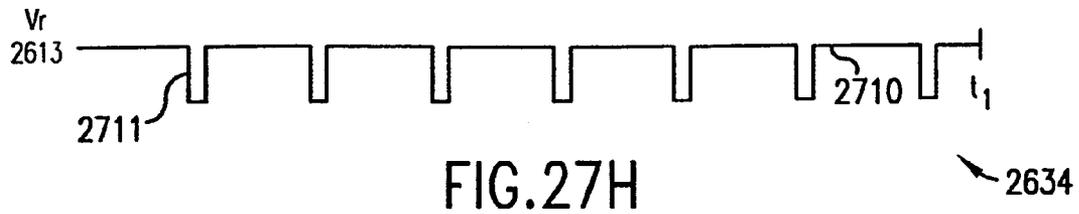


FIG. 27H

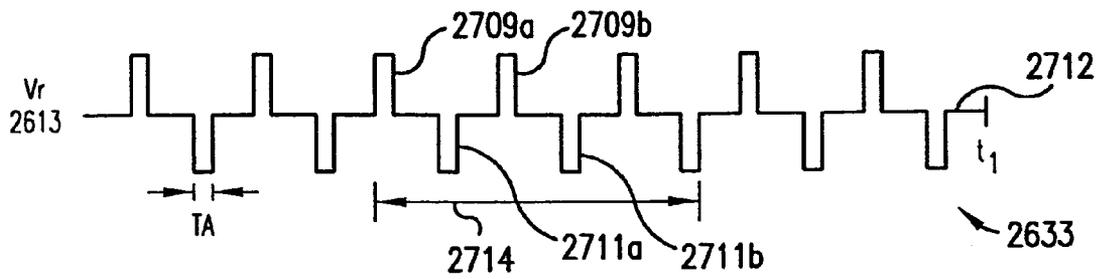


FIG. 27I

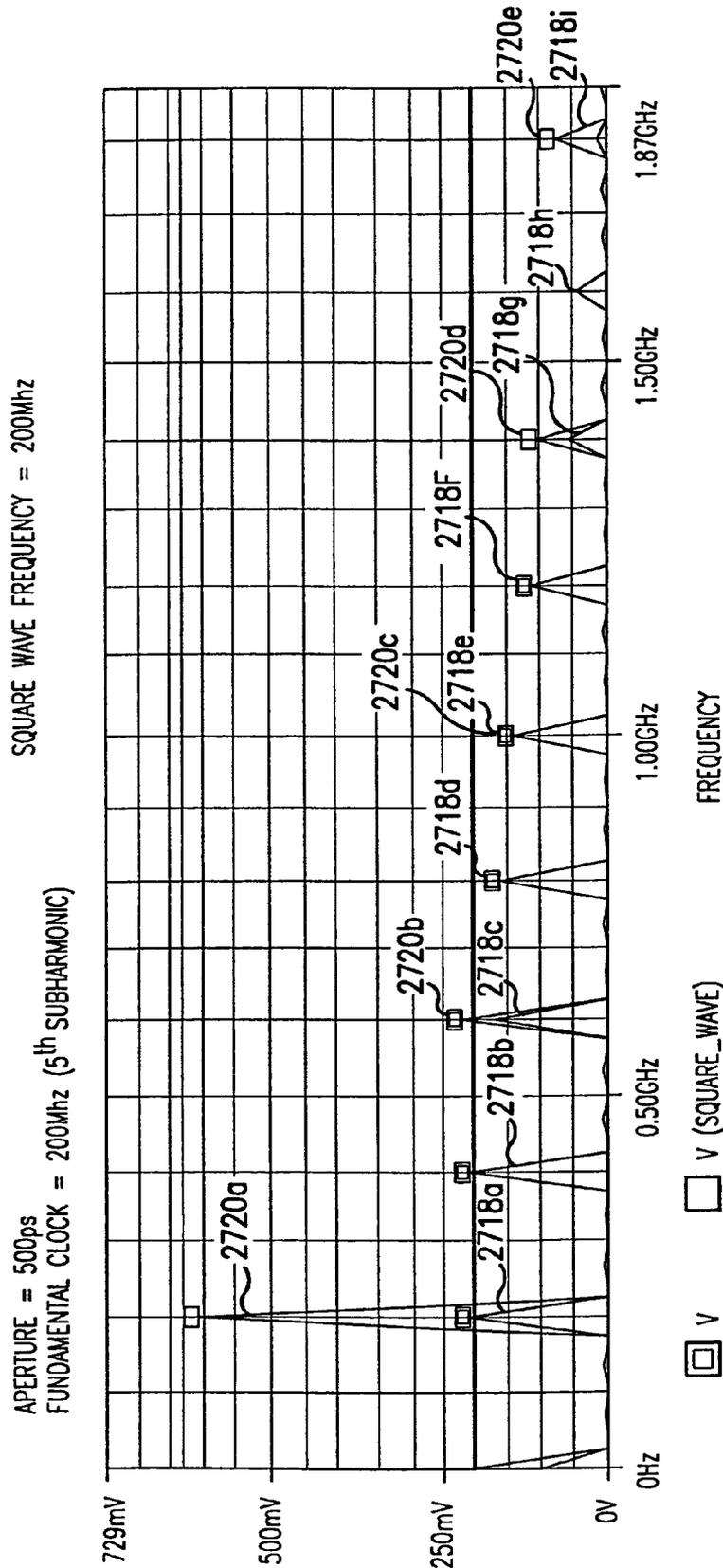
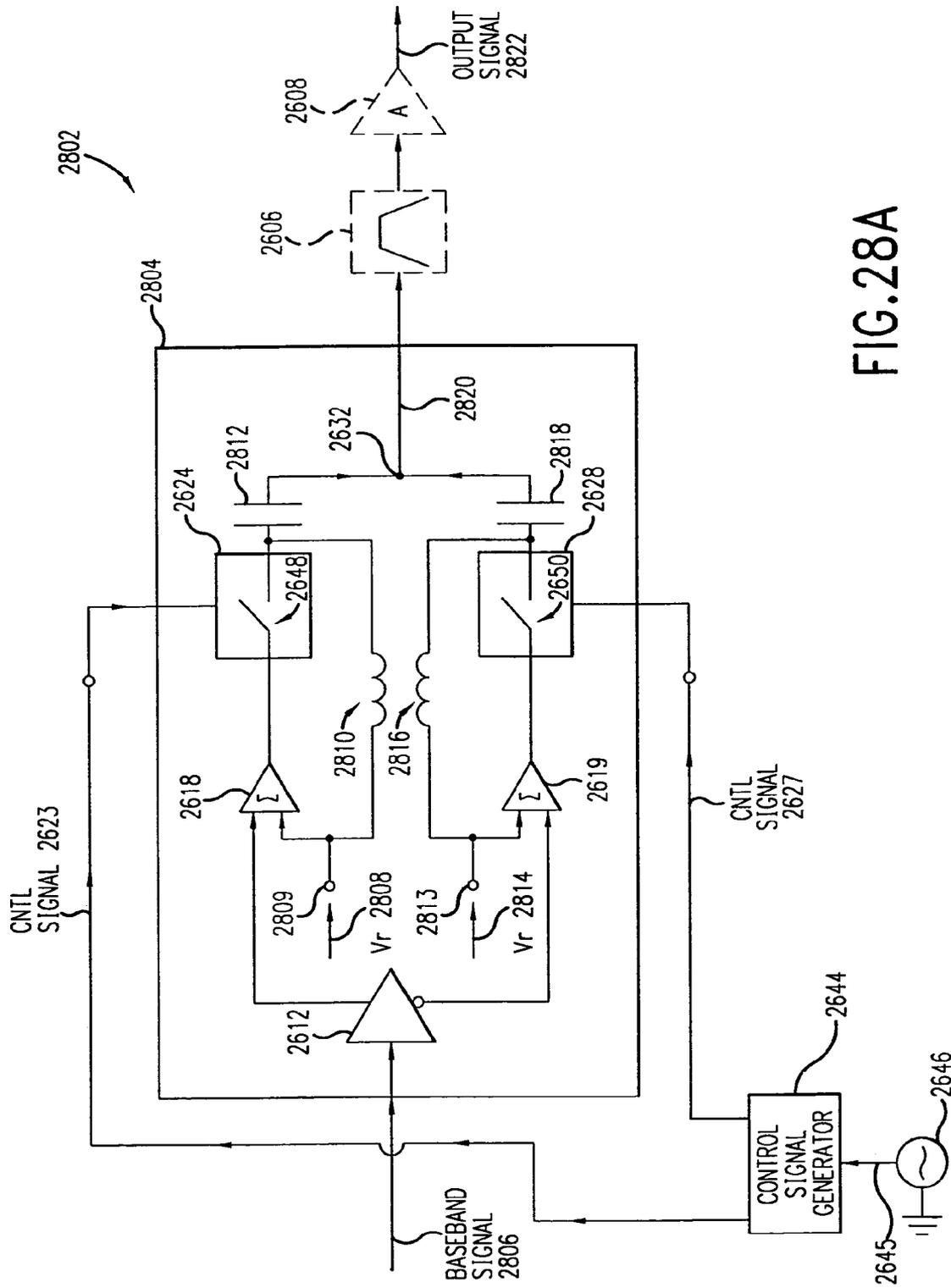


FIG.27J



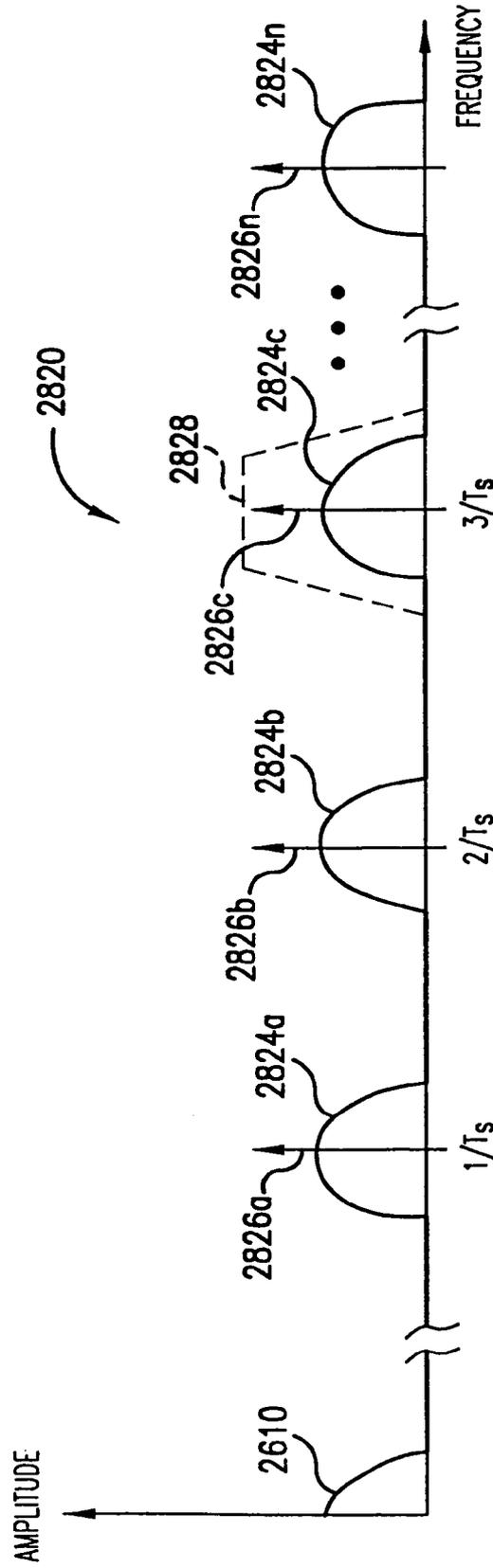


FIG.28B

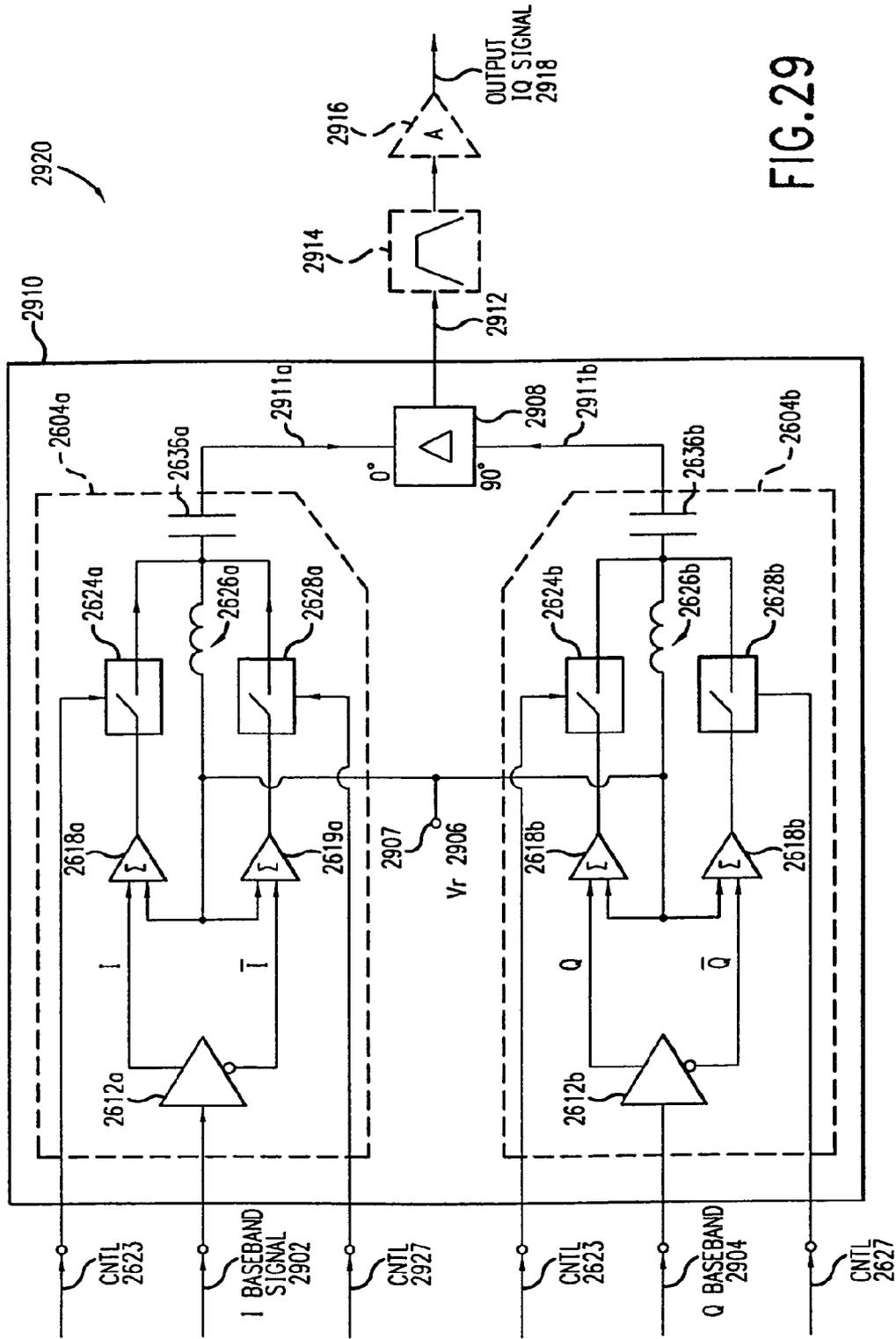


FIG. 29

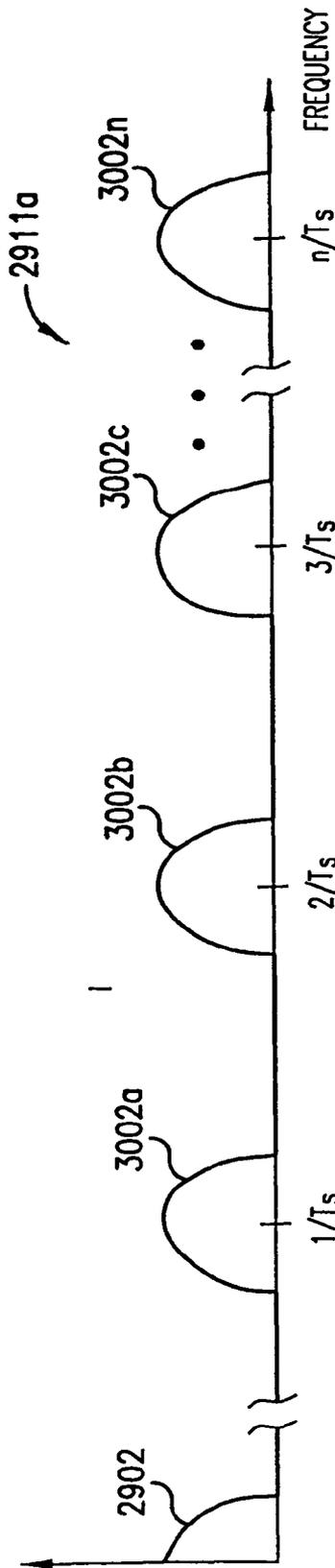


FIG. 30A

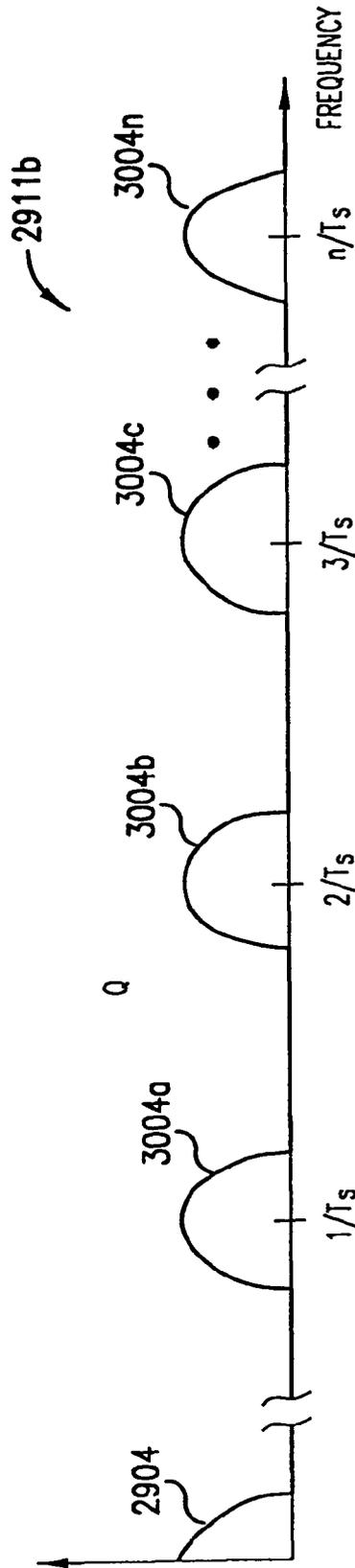


FIG. 30B

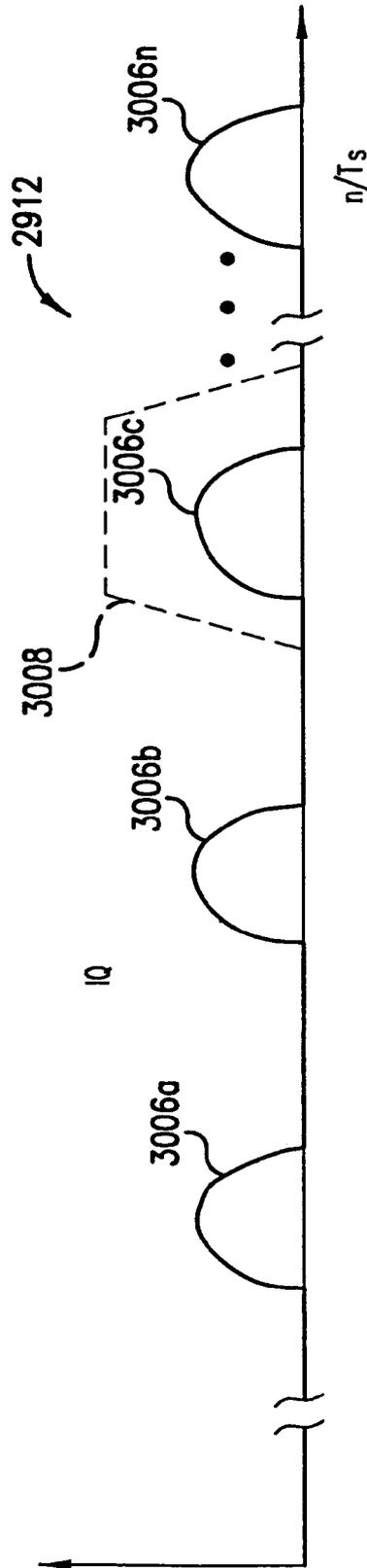


FIG.30C

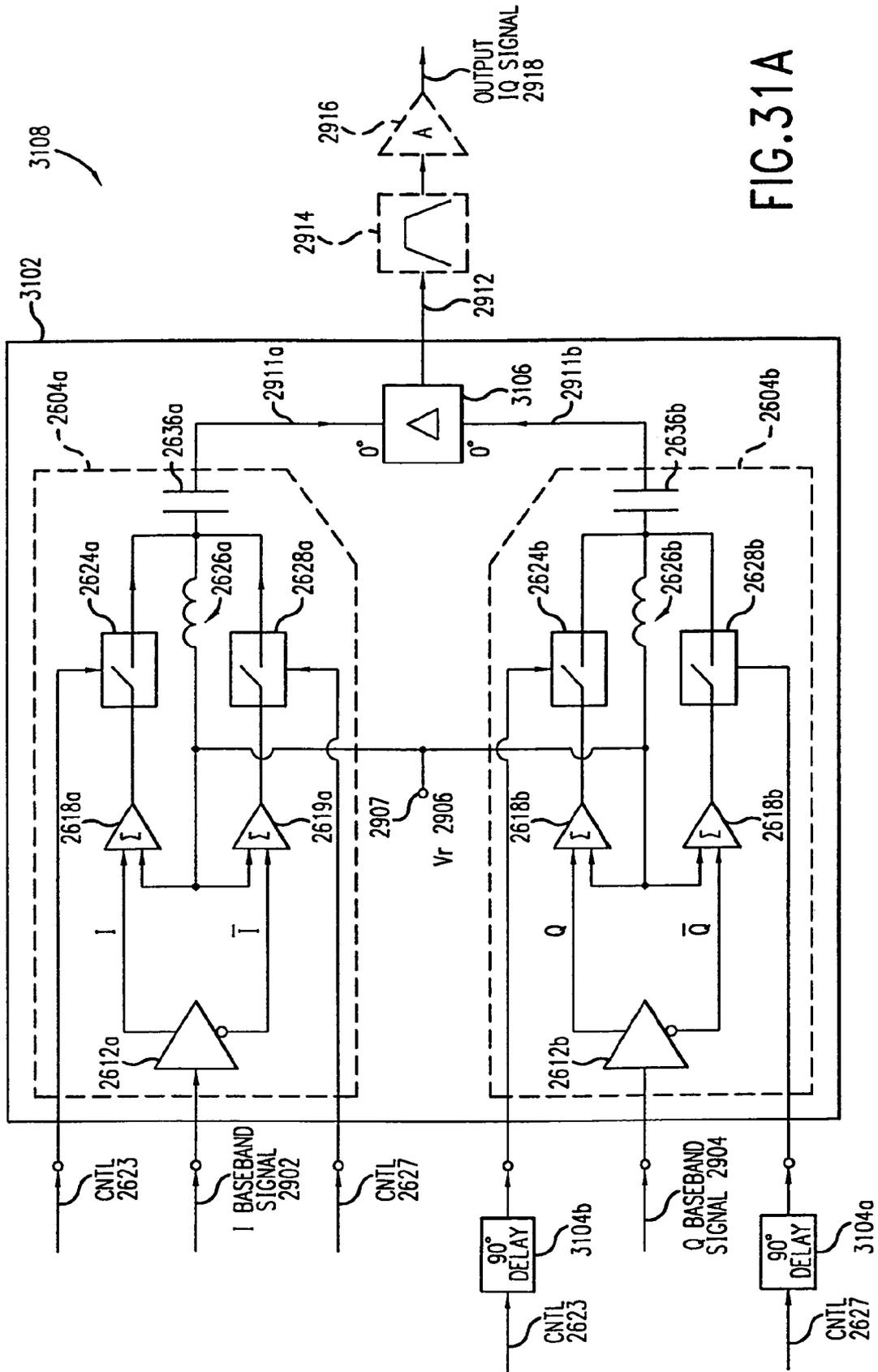


FIG. 31A

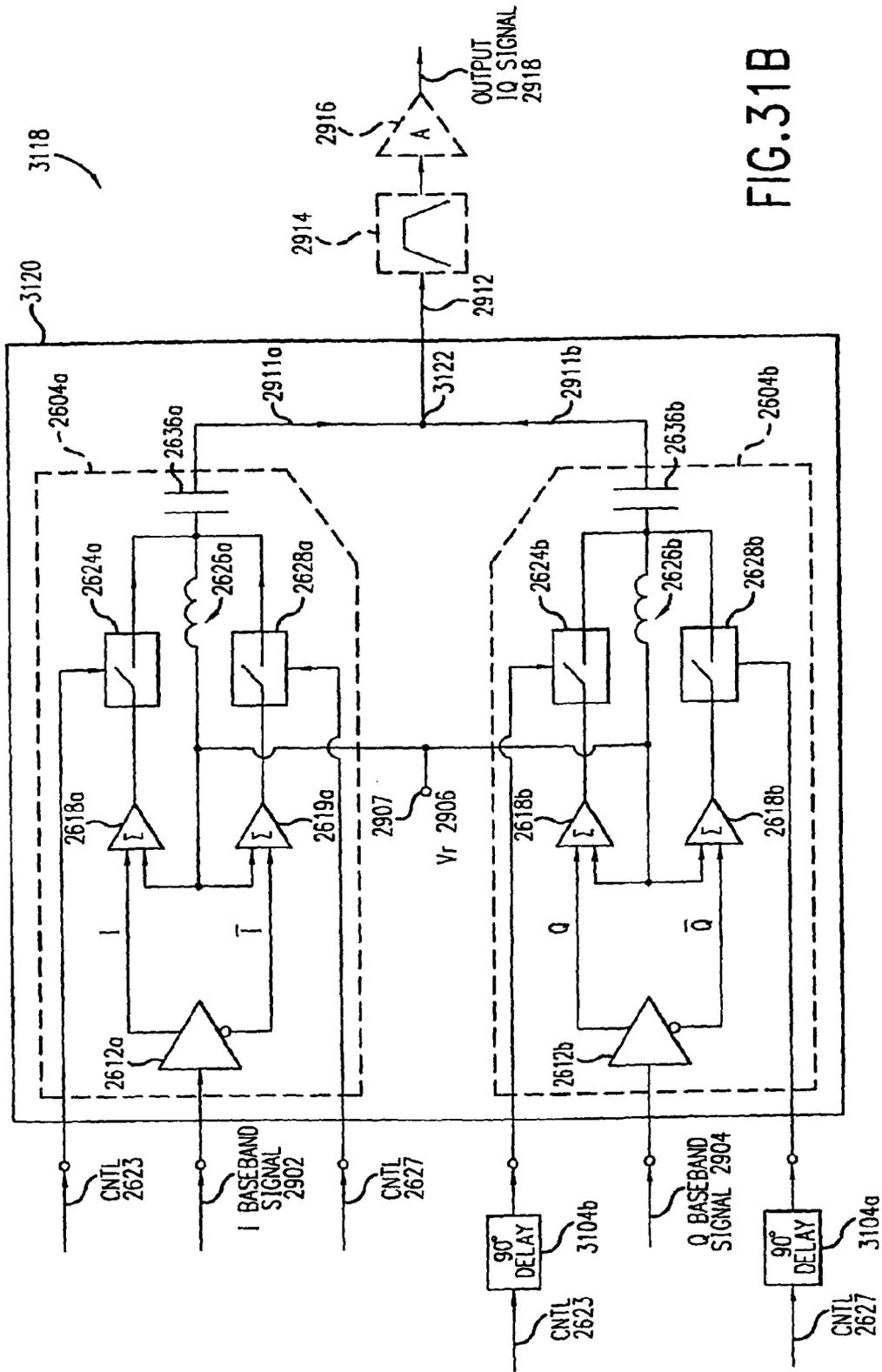


FIG. 311B

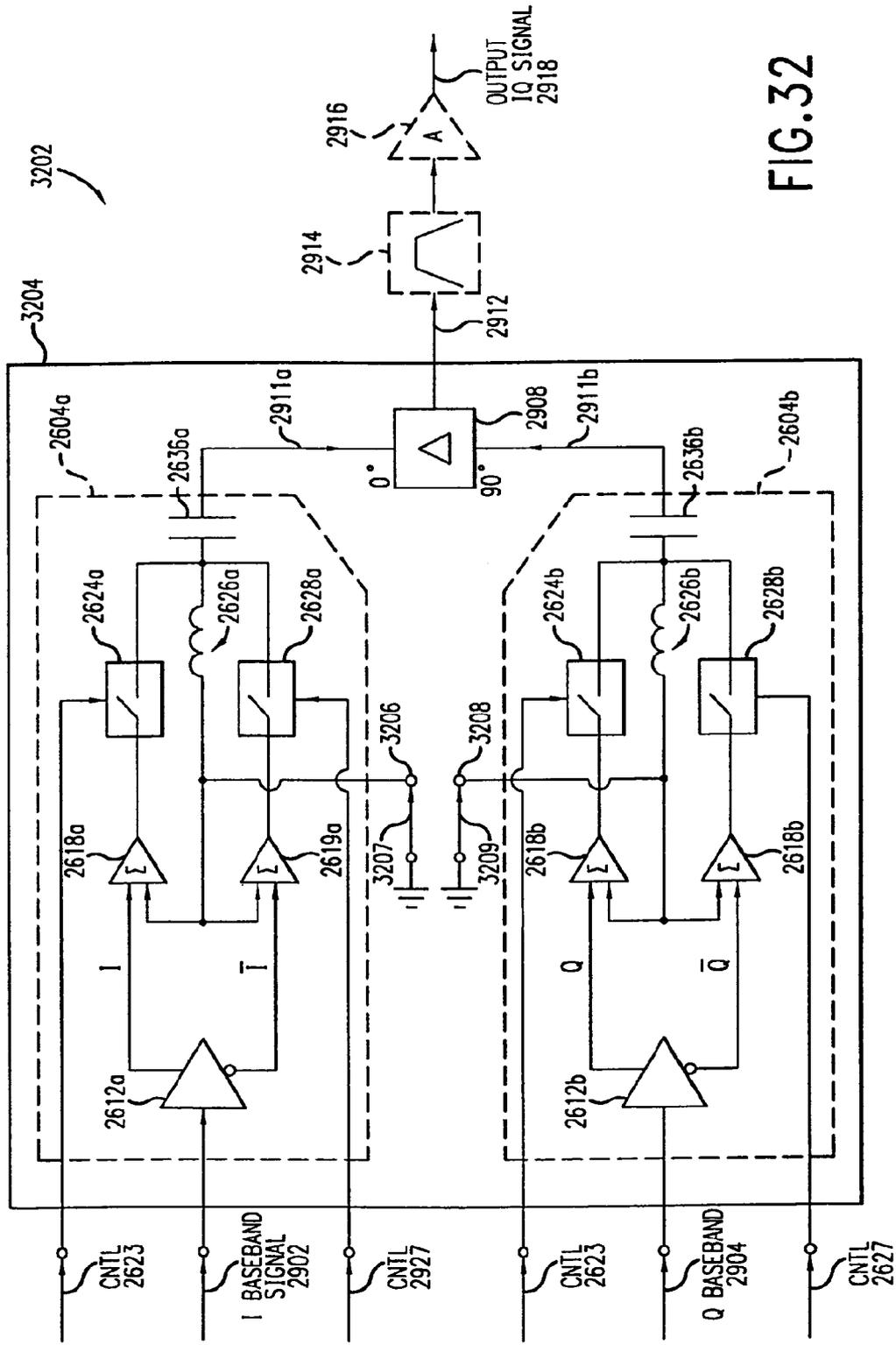


FIG. 32

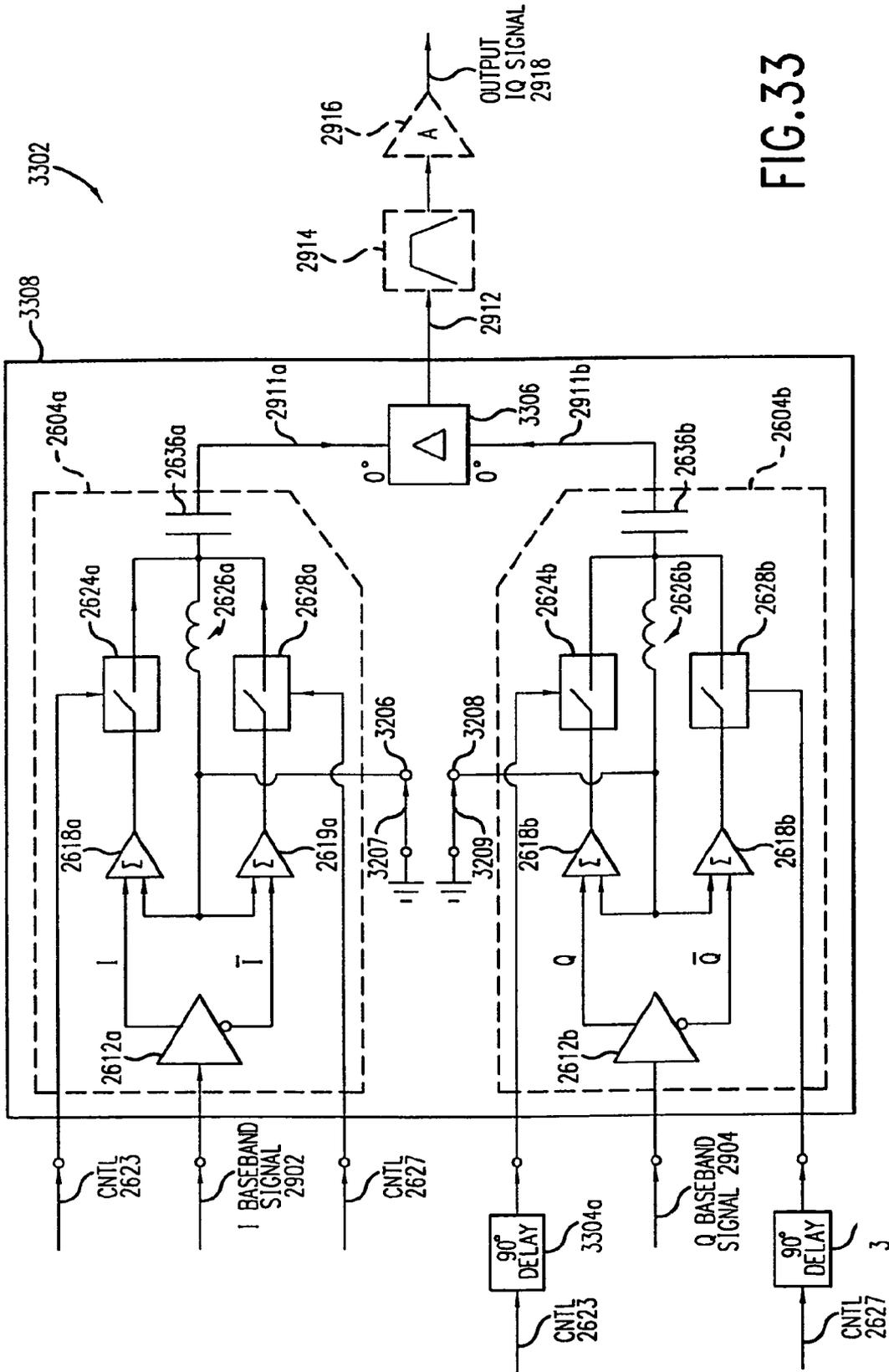


FIG. 33

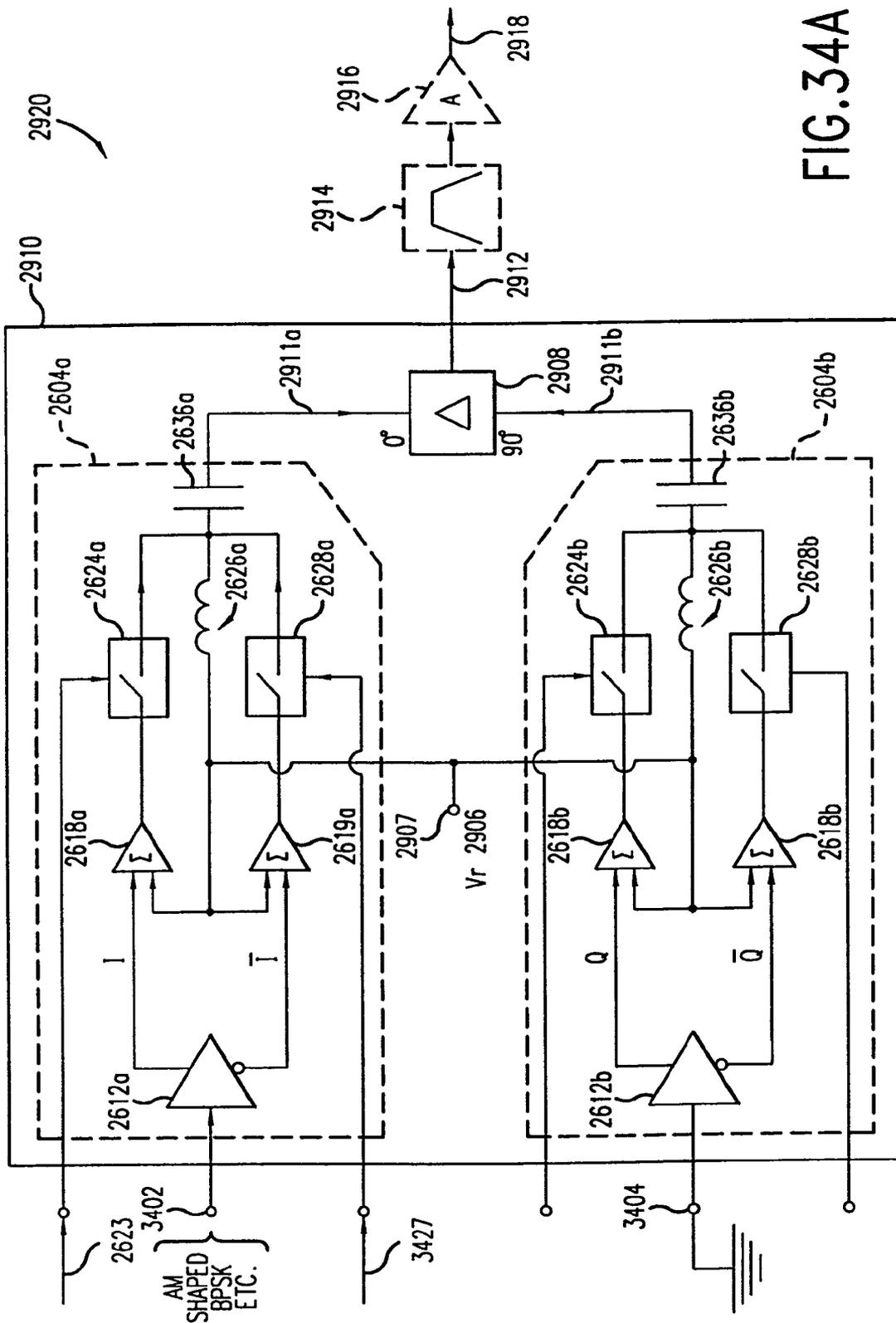


FIG. 34A

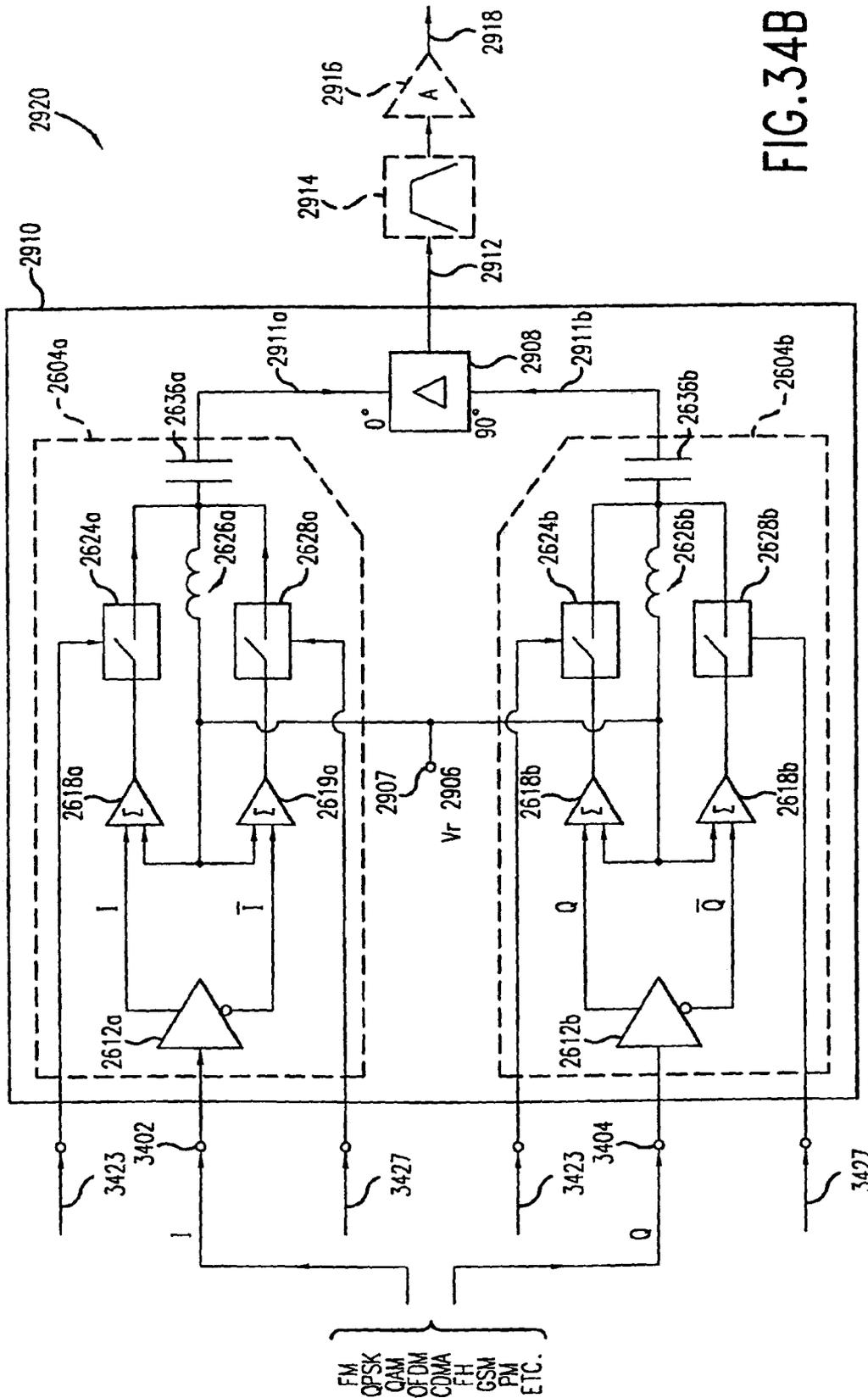


FIG. 34B

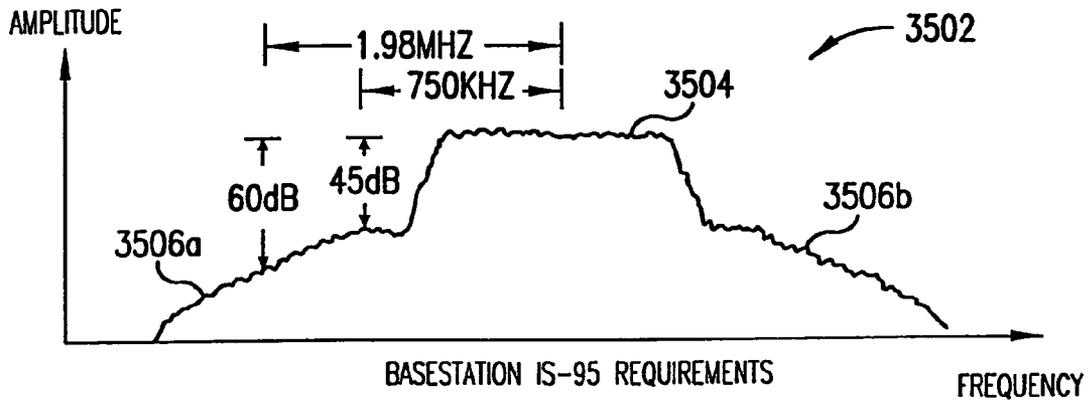


FIG.35A

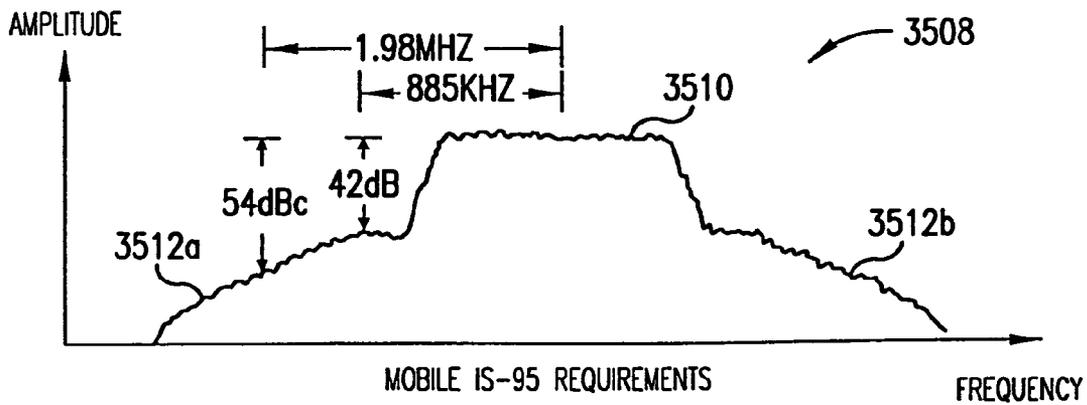
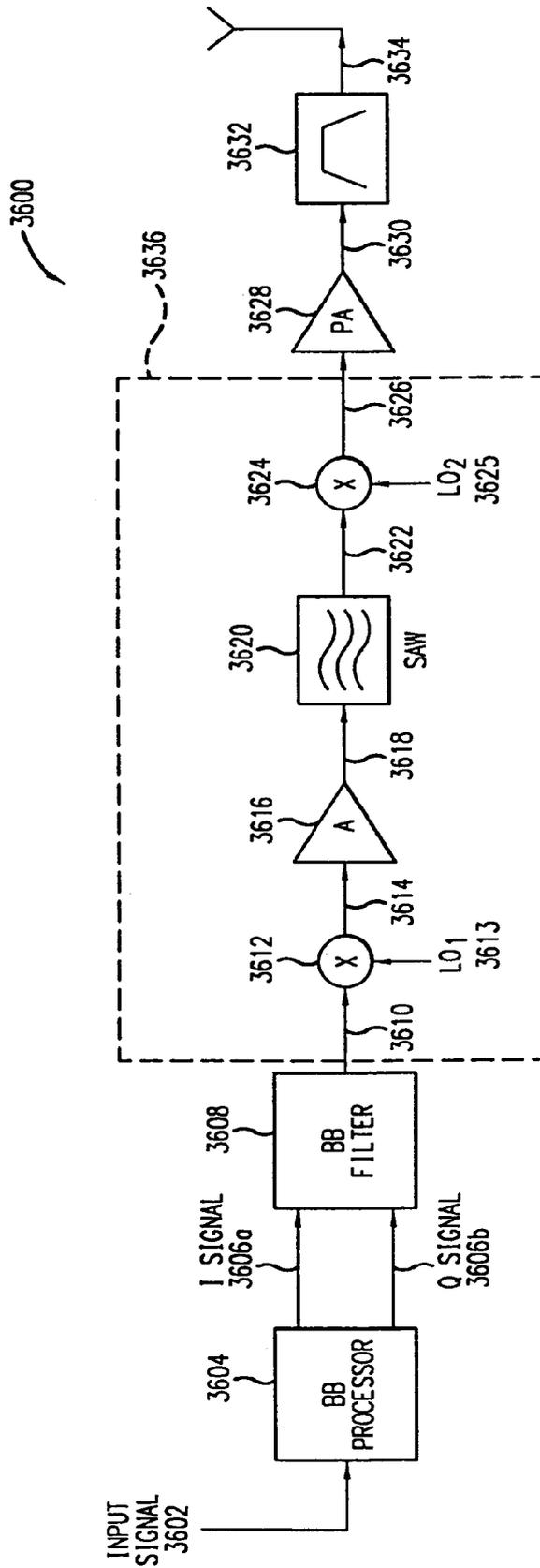


FIG.35B



CONVENTIONAL TRANSMITTER

FIG.36

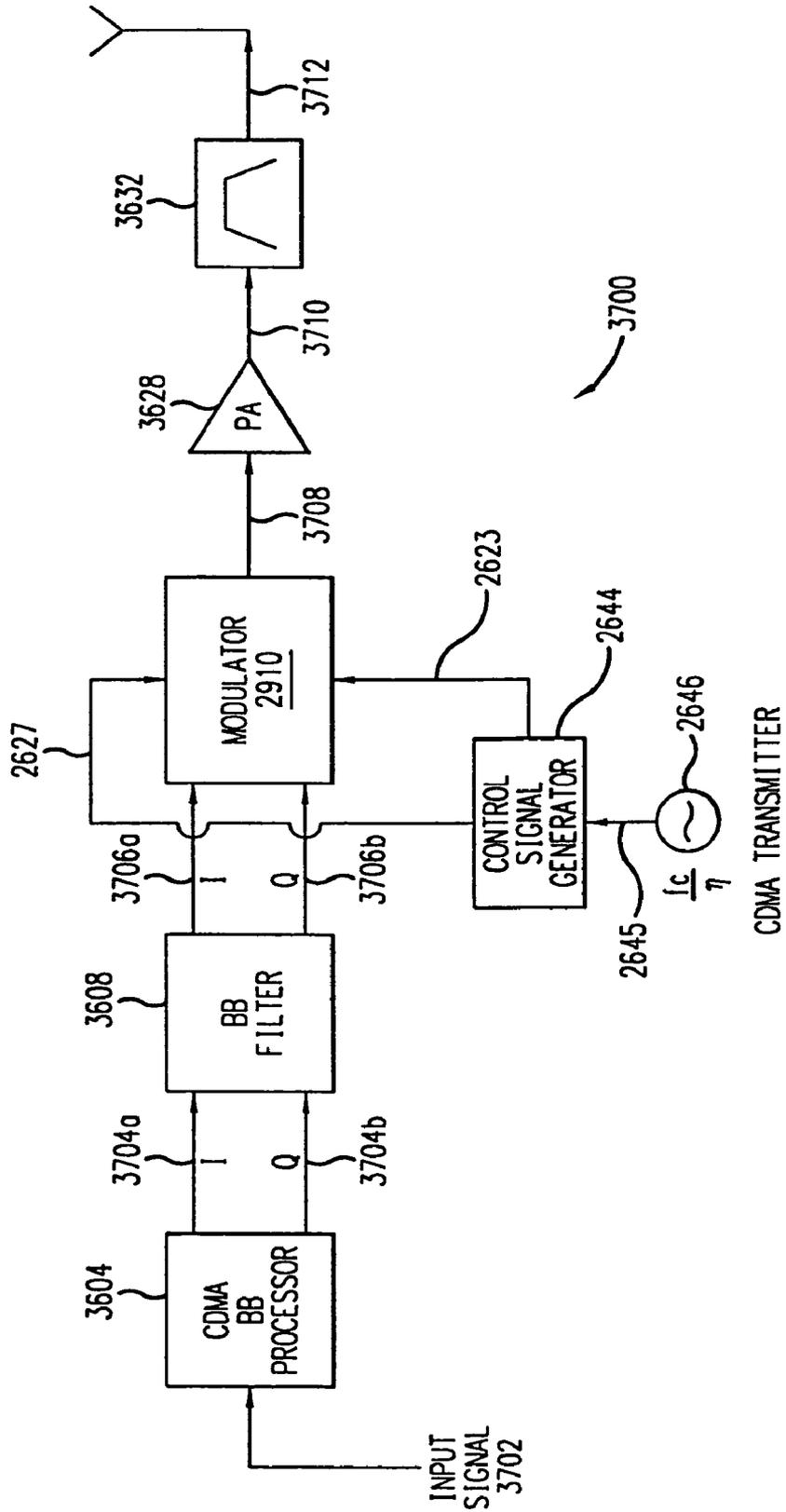


FIG.37A

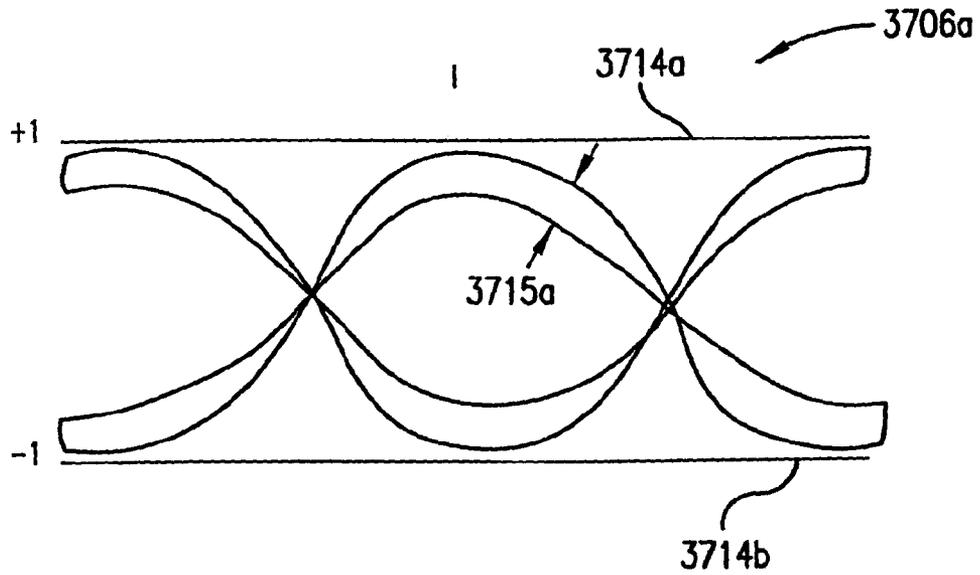


FIG.37B

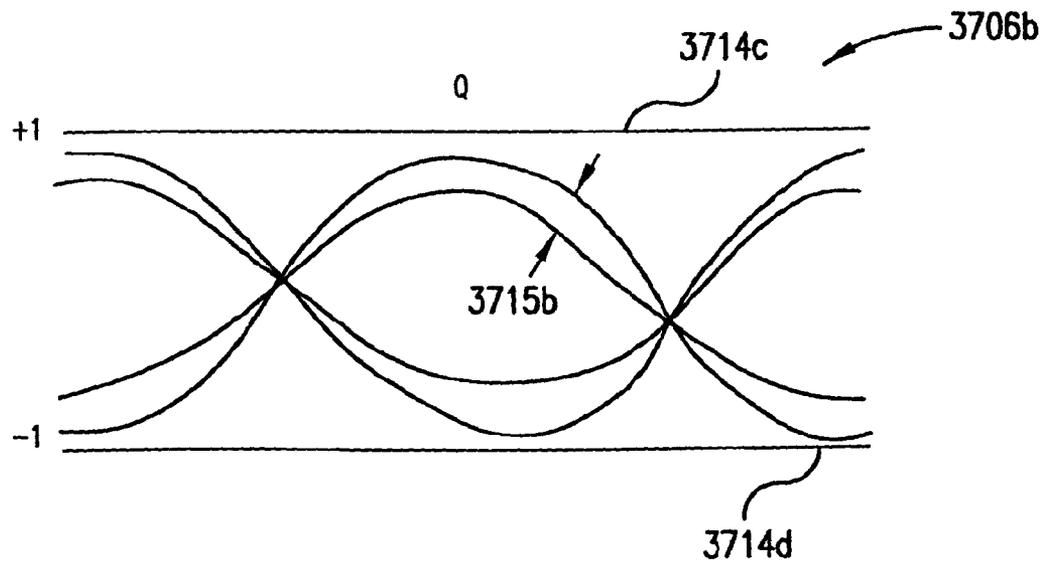


FIG.37C

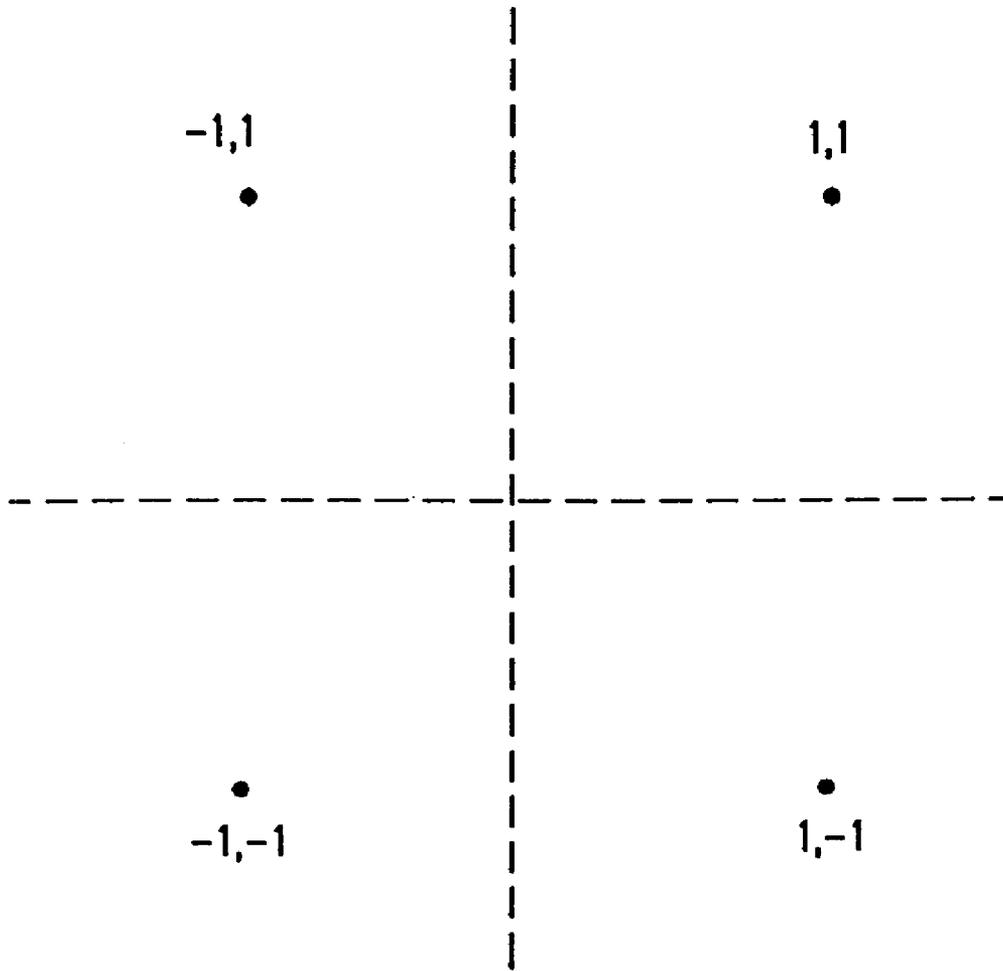


FIG.37D

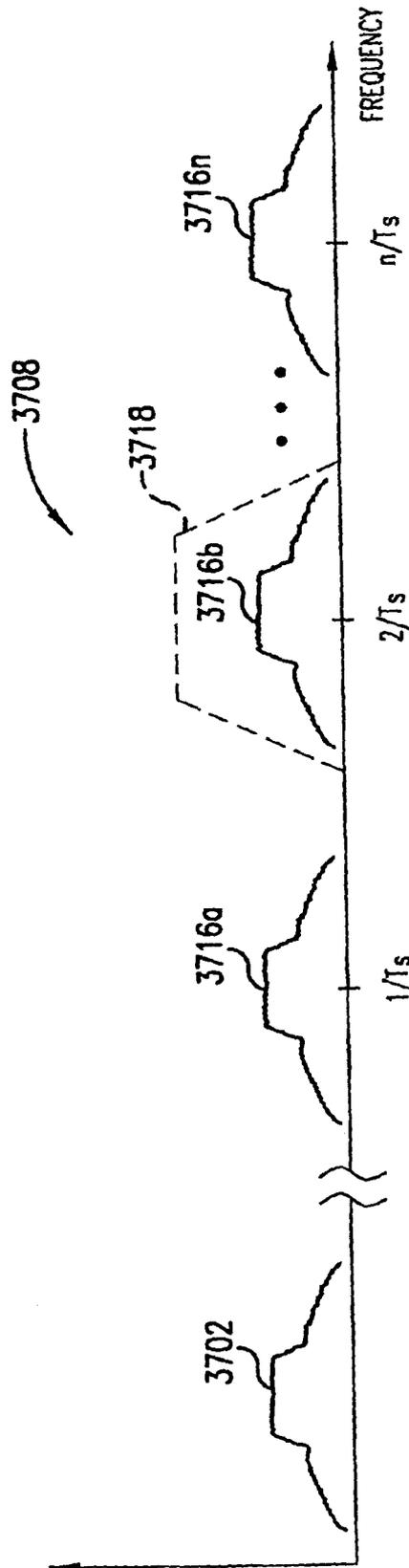


FIG. 37E

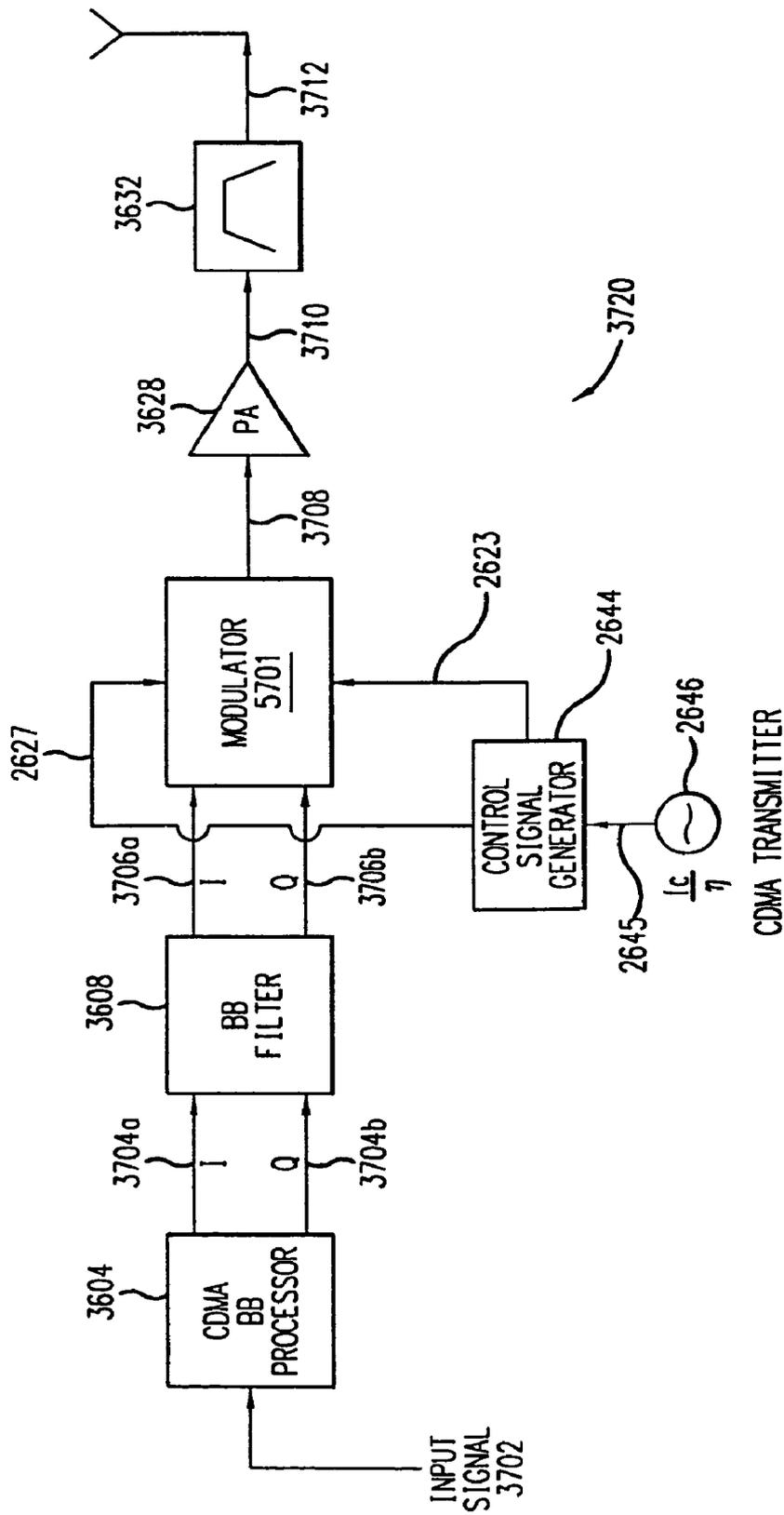
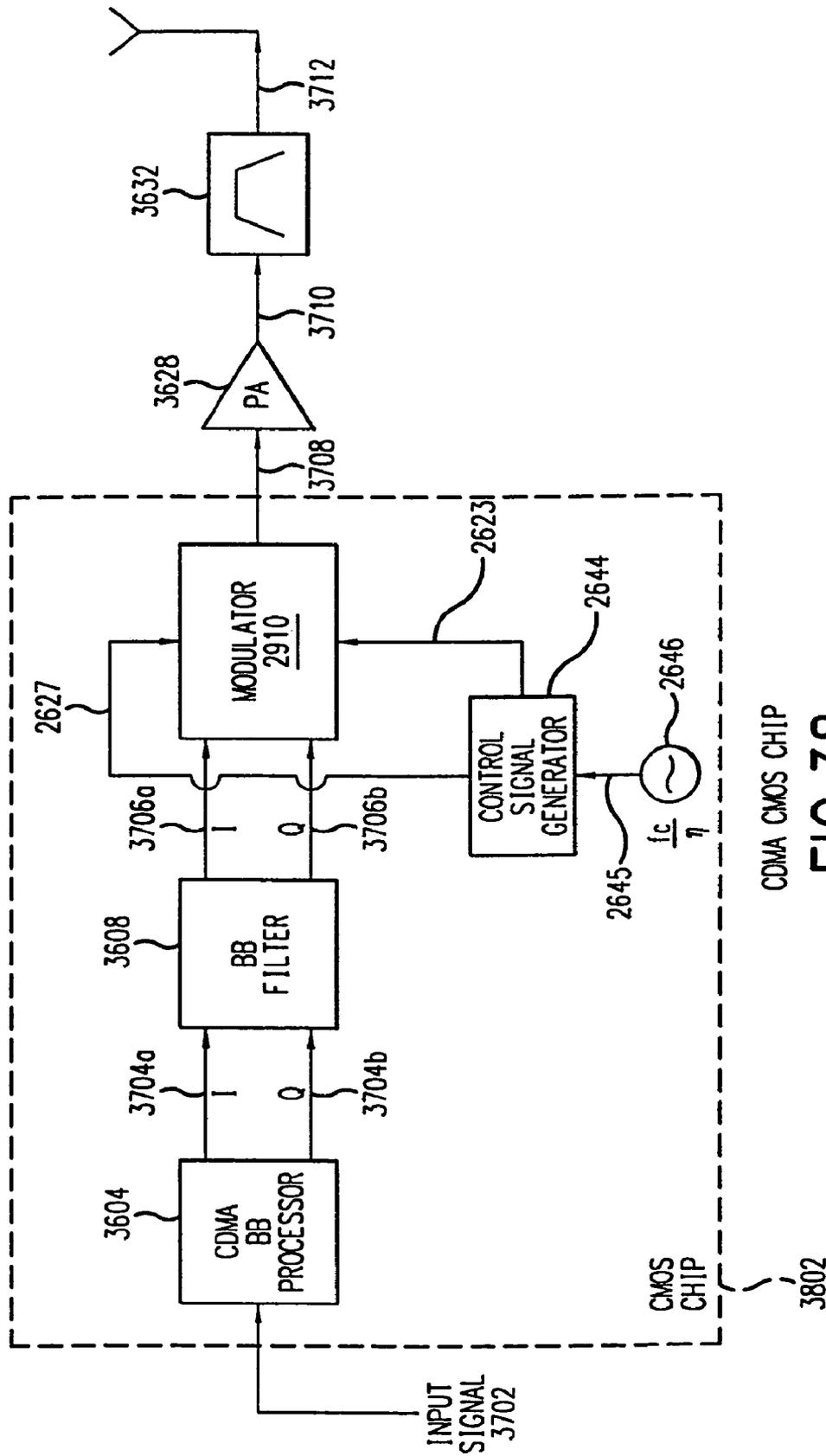


FIG.37F



CDMA CMOS CHIP

FIG.38

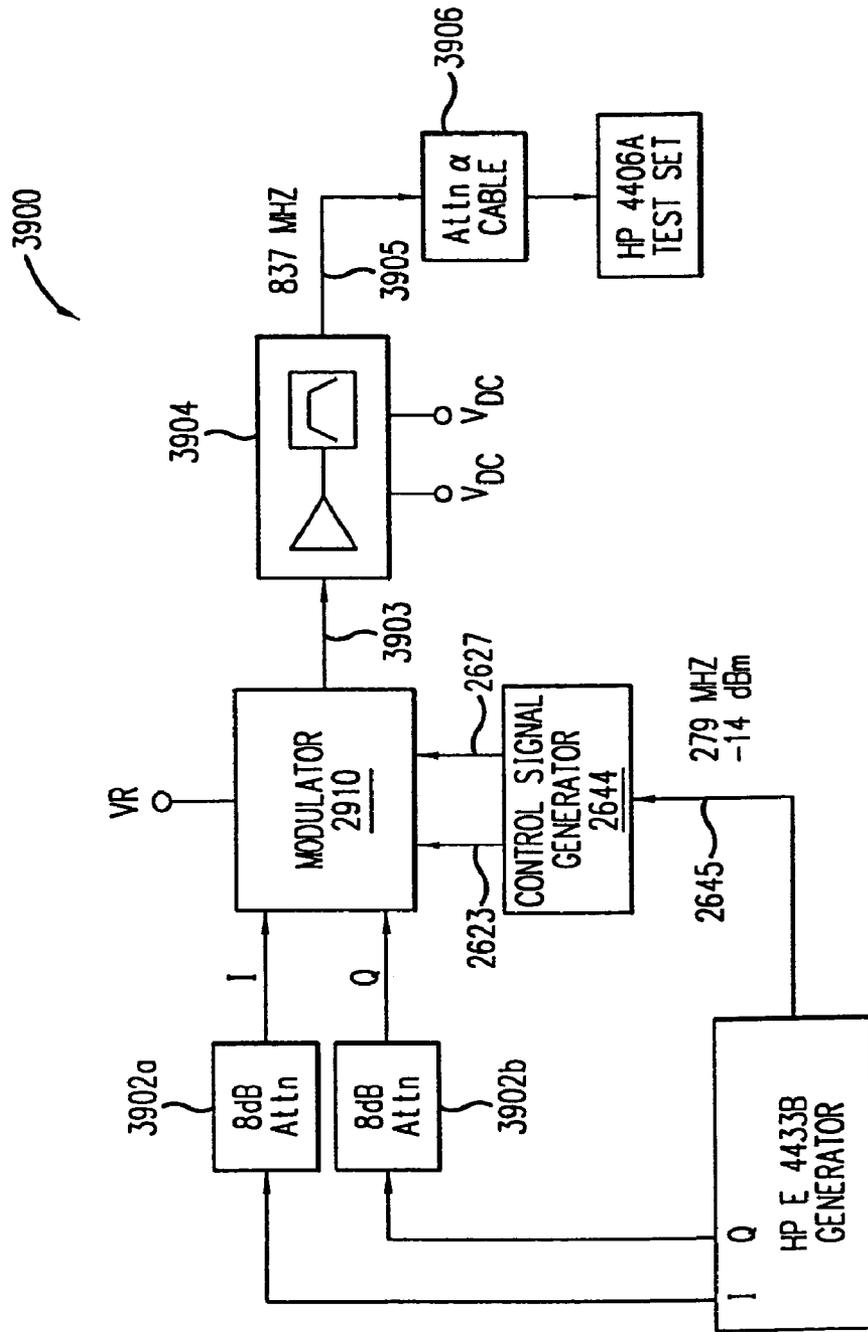


FIG. 39

4002

BASE STATION

RHO	0.9970
EVM	5.51%
PHASE ERROR	1.80°
MAGNITUDE ERROR	4.53%
CARRIER INSERTION	-37.91 dB
PA POWER OUT	28.06 dBm

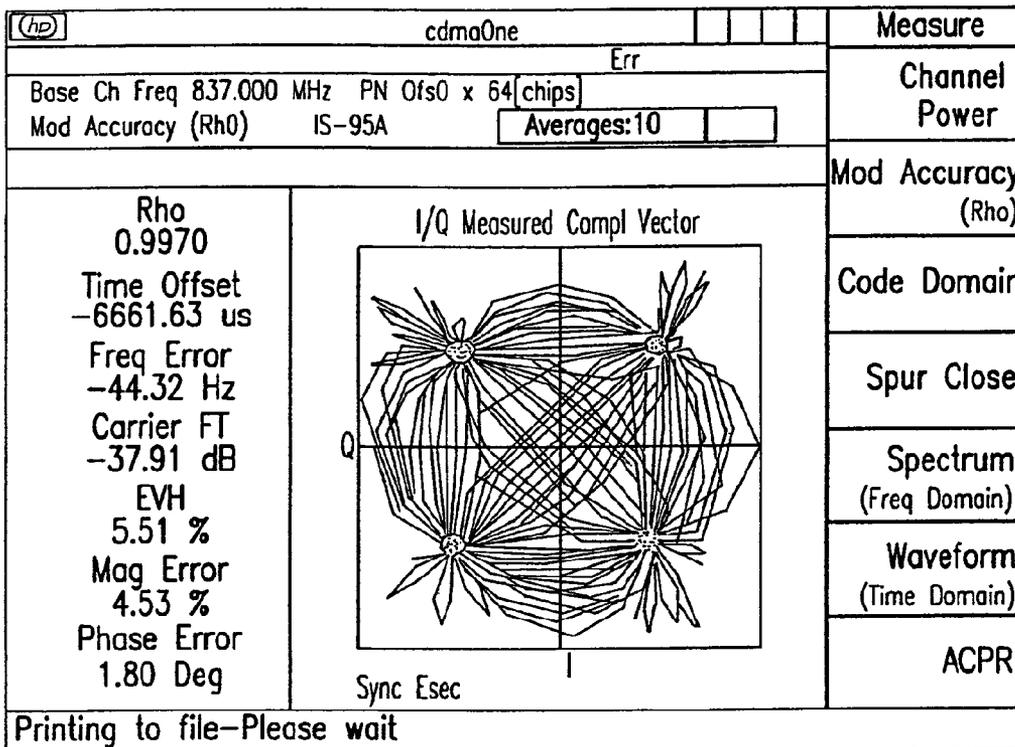
FIG.40

FREQUENCY (MHz) (MOBILE STATION)

4102

	LOW	MIDDLE	HIGH
RHO	0.9892	0.9969	0.9892
EVM	10.39%	5.54%	10.39%
PHASE ERROR	4.47°	2.24°	4.08°
MAGNITUDE ERROR	6.84%	4.21%	8.27%
CARRIER INSERTION	-40.15 dB	-44.58 dB	-35.27 dB
PA POWER OUT	27.36 dBm	28.11 dBm	27.55 dBm

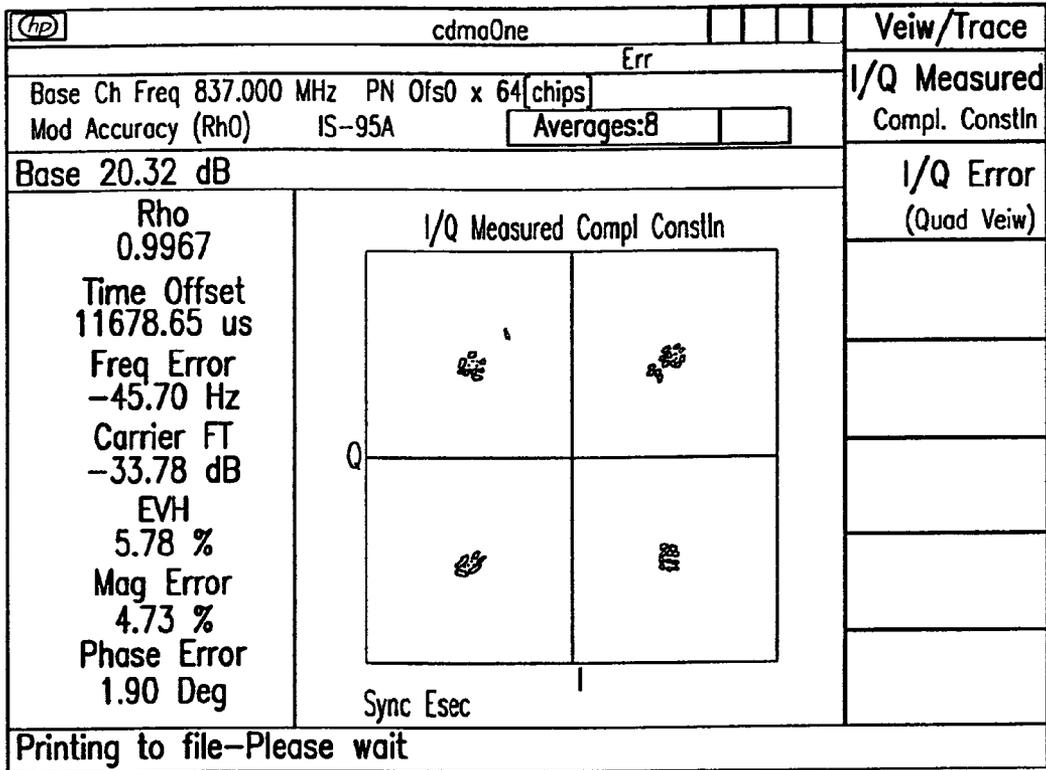
FIG.41



BASE STATION CONSTELLATION FOR PILOT CHANNEL TEST

FIG.42

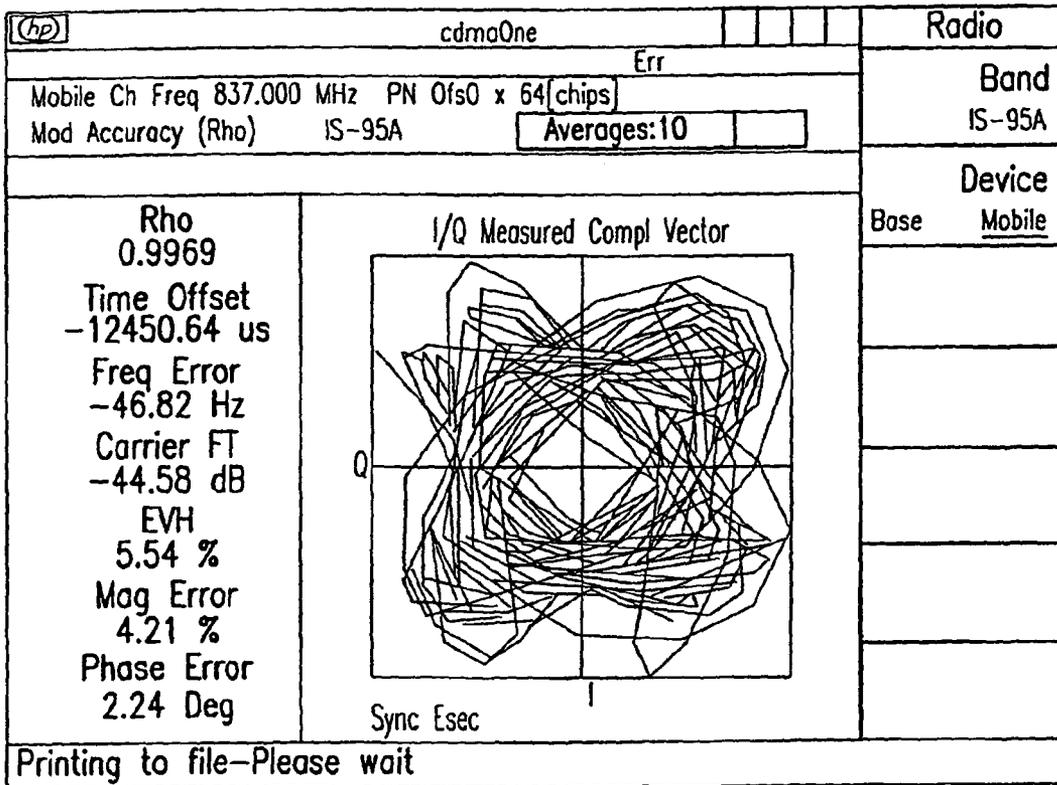
4202



BASE STATION SAMPLED CONSTELLATION

FIG.43

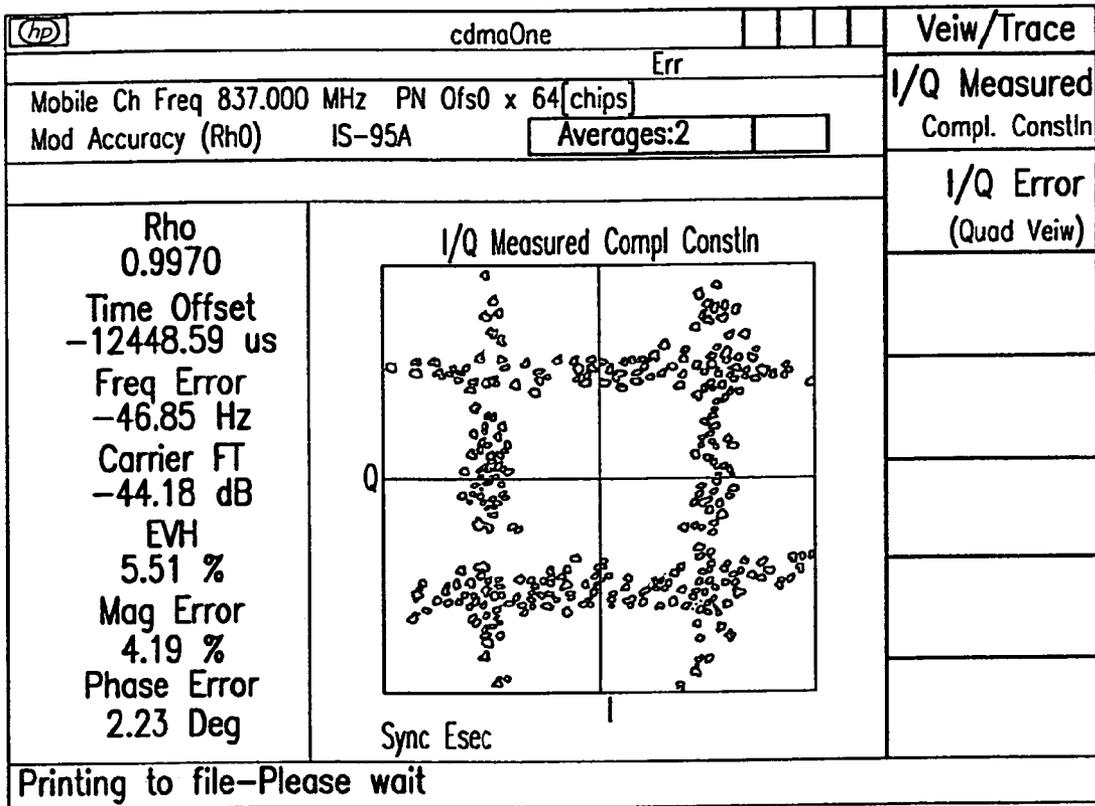
4302



MOBILE STATION CONSTELLATION FOR ACCESS CHANNEL TEST

FIG.44

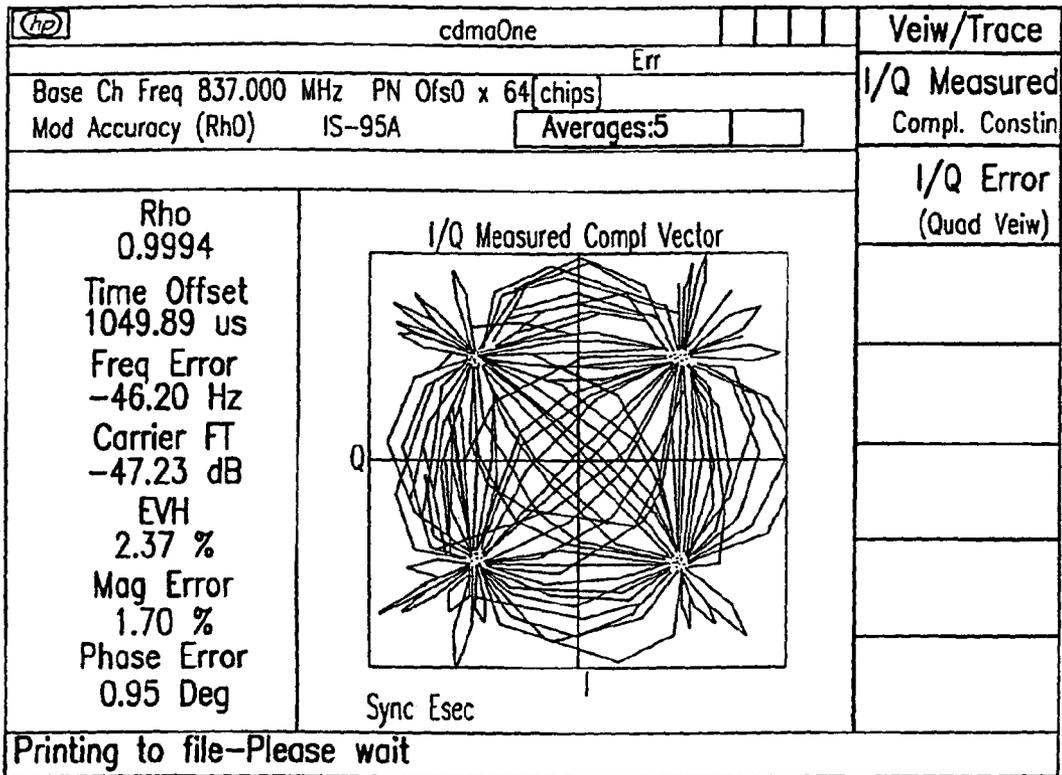
4402



MOBILE STATION SAMPLED CONSTELLATION

FIG.45

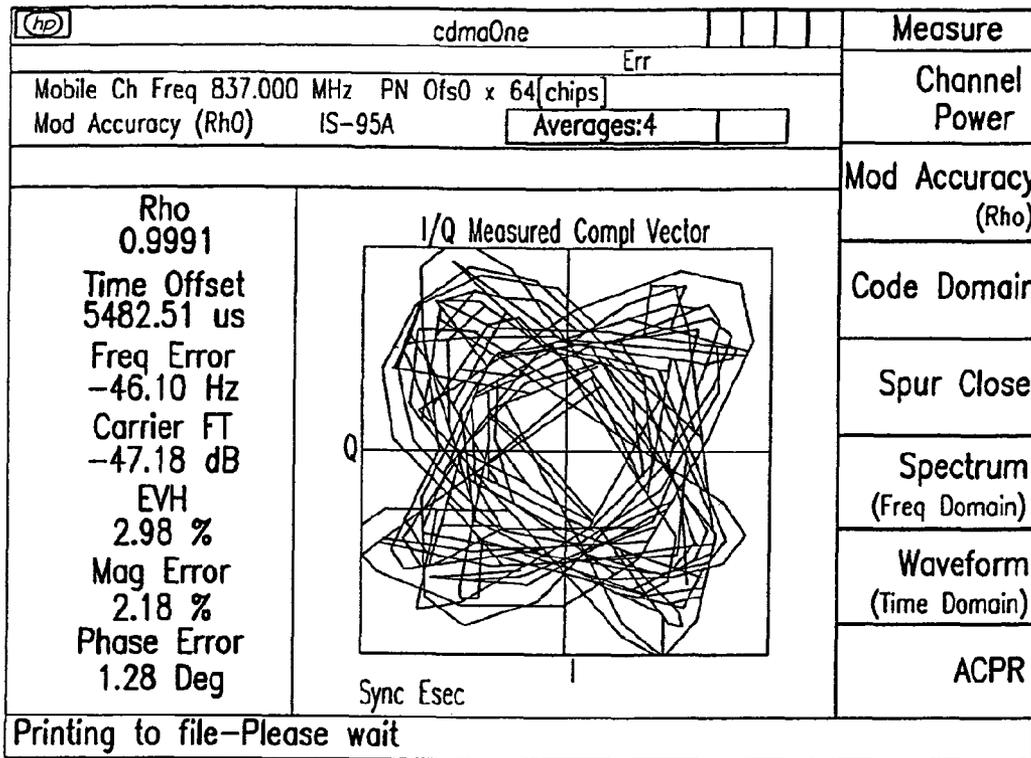
4502



BASE STATION CONSTELLATION USING ONLY H/P TEST EQUIPMENT

FIG.46

4602



MOBILE CONSTELLATION USING ONLY H/P TEST EQUIPMENT

FIG.47

4702

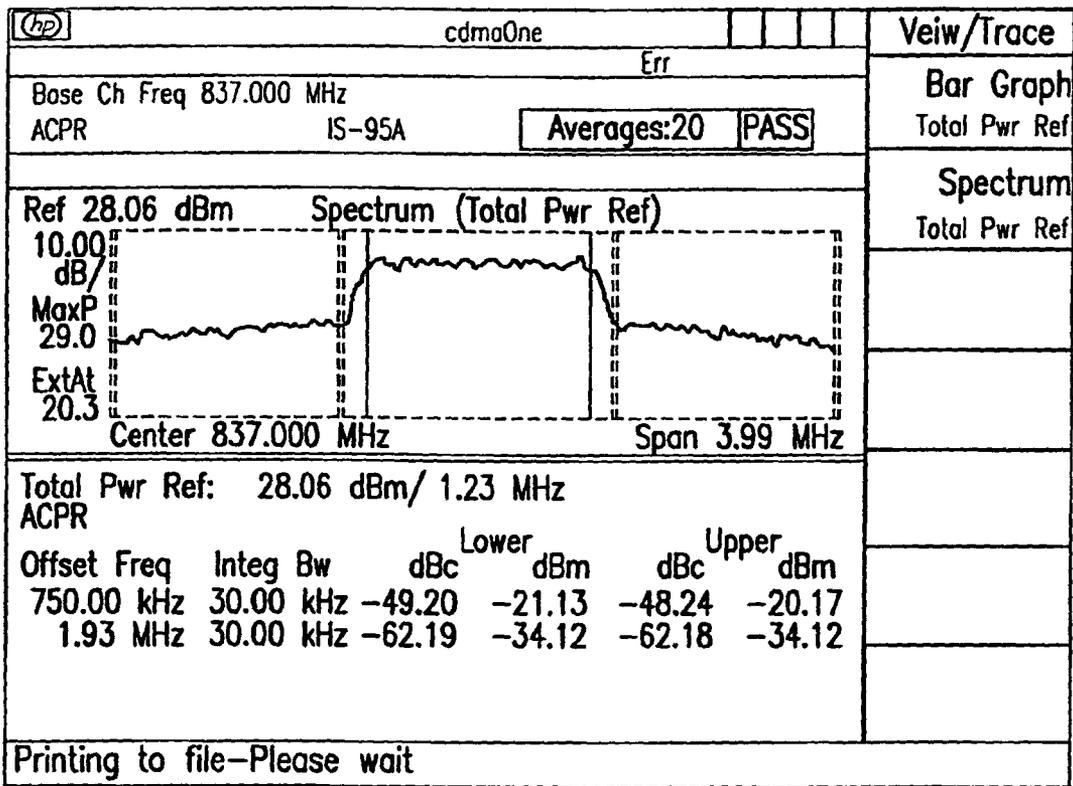
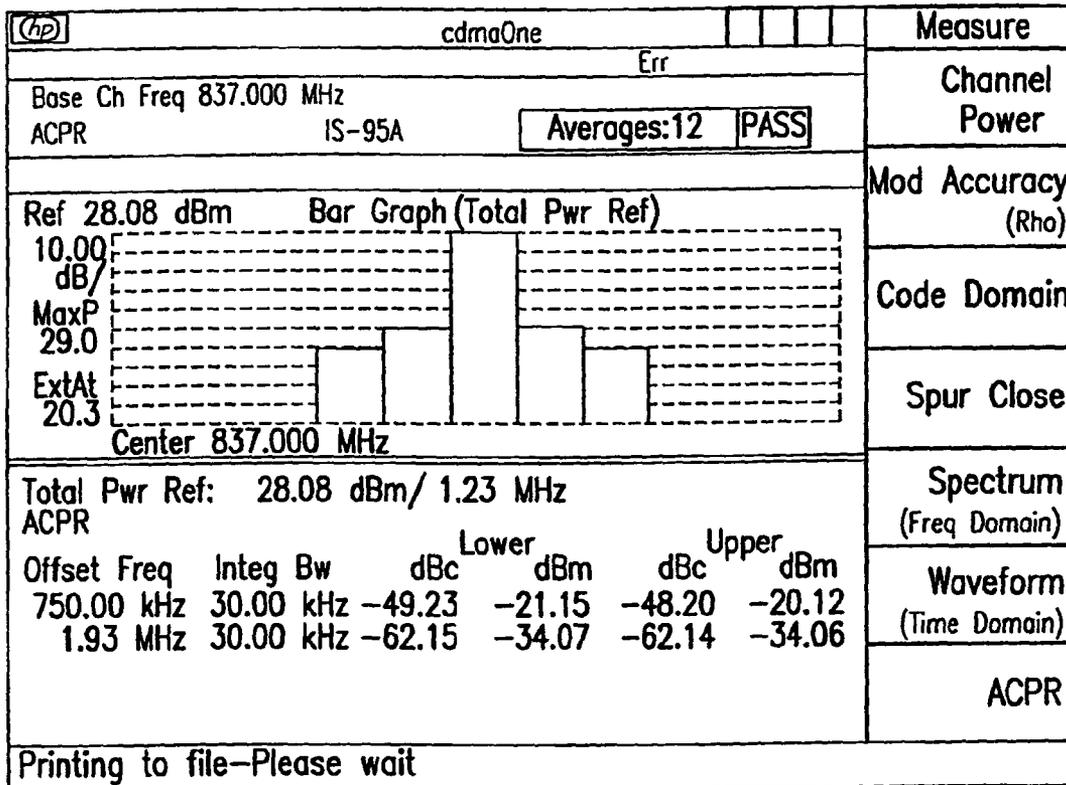


FIG. 48

4802



BASE STATION SPECTRAL RESPONSE WITH MASK

FIG.49

4902

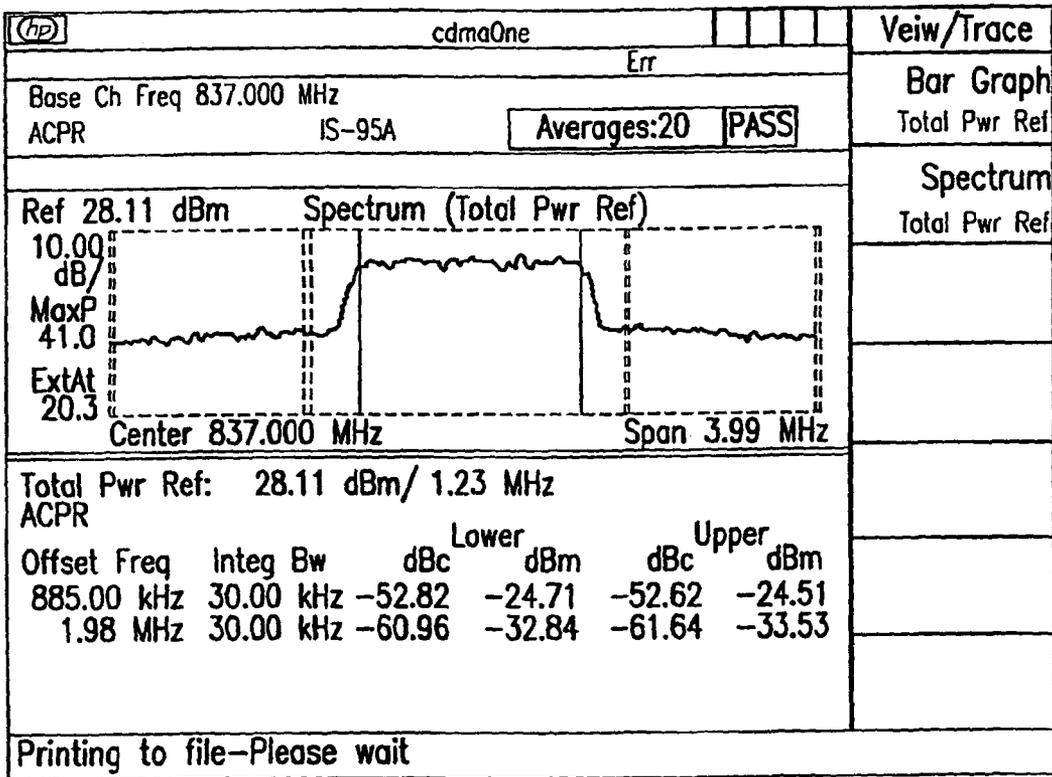
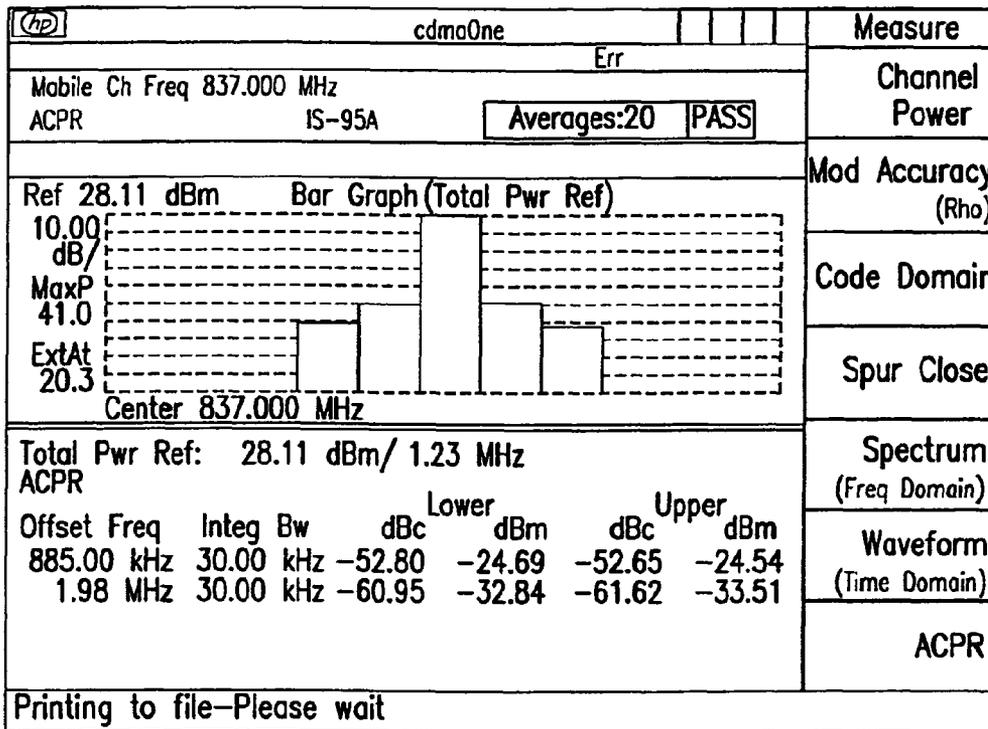


FIG. 50

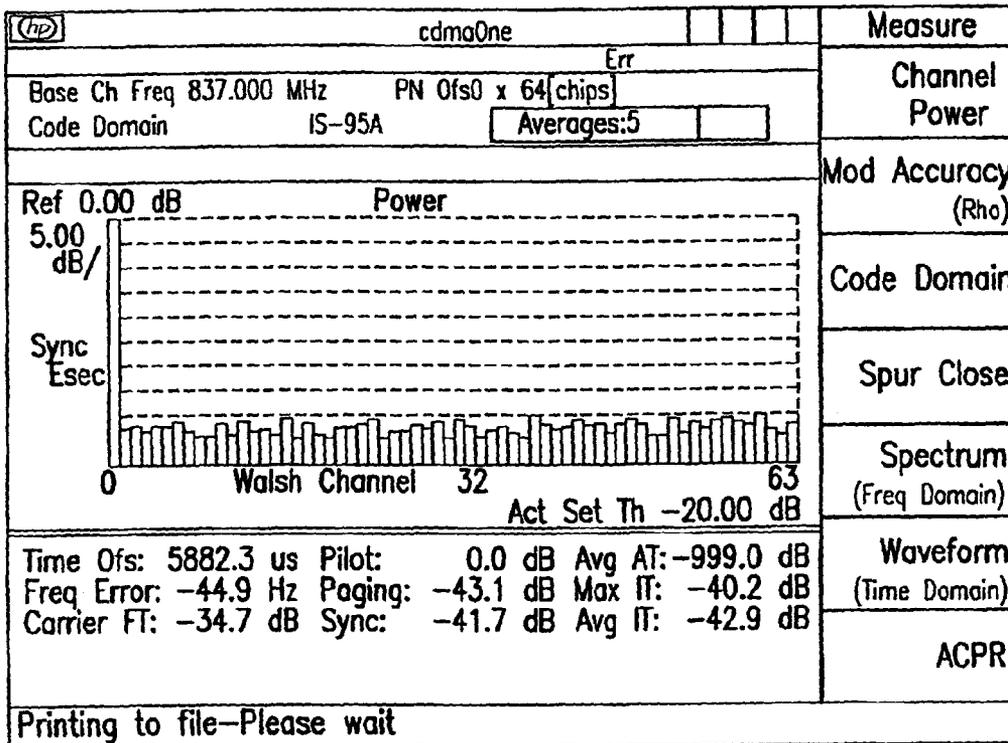
5002



MOBILE STATION SPECTRAL RESPONSE WITH MASK

FIG.51

5102



CDMA CROSSTALK

FIG.52A

5202

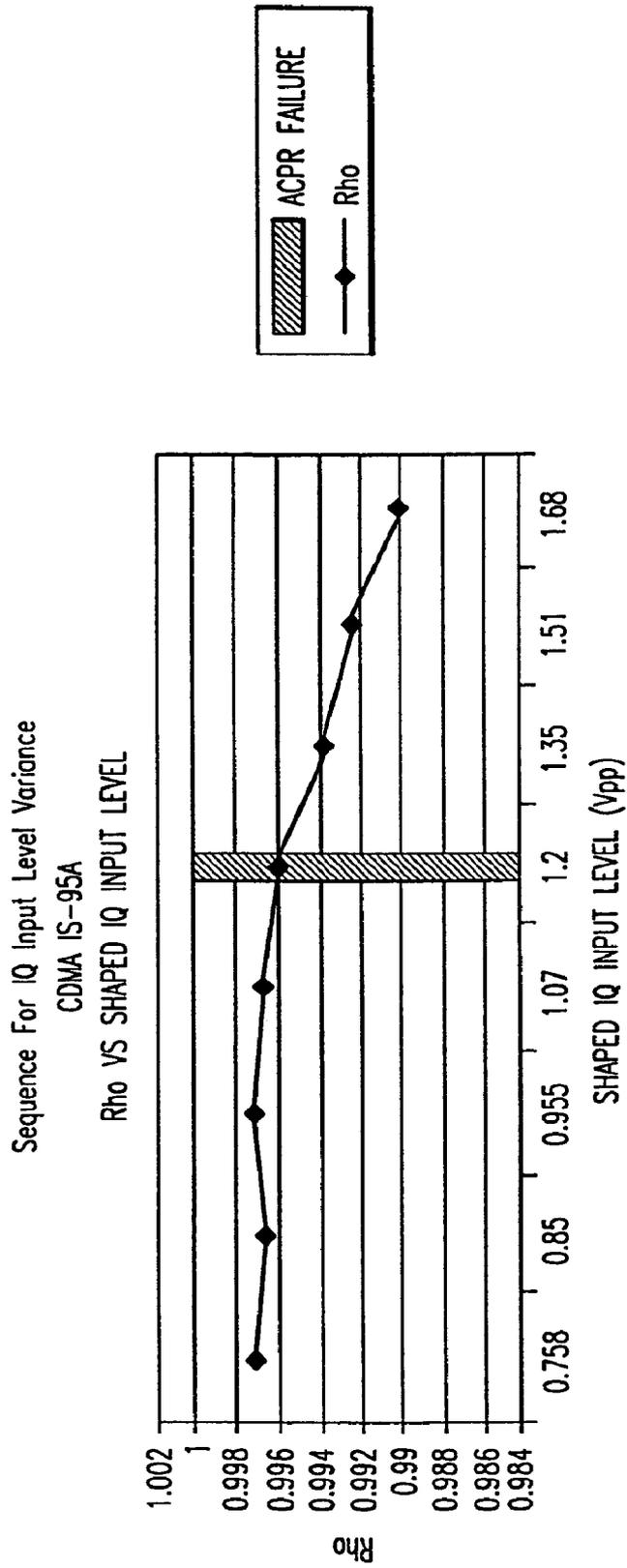


FIG. 52B

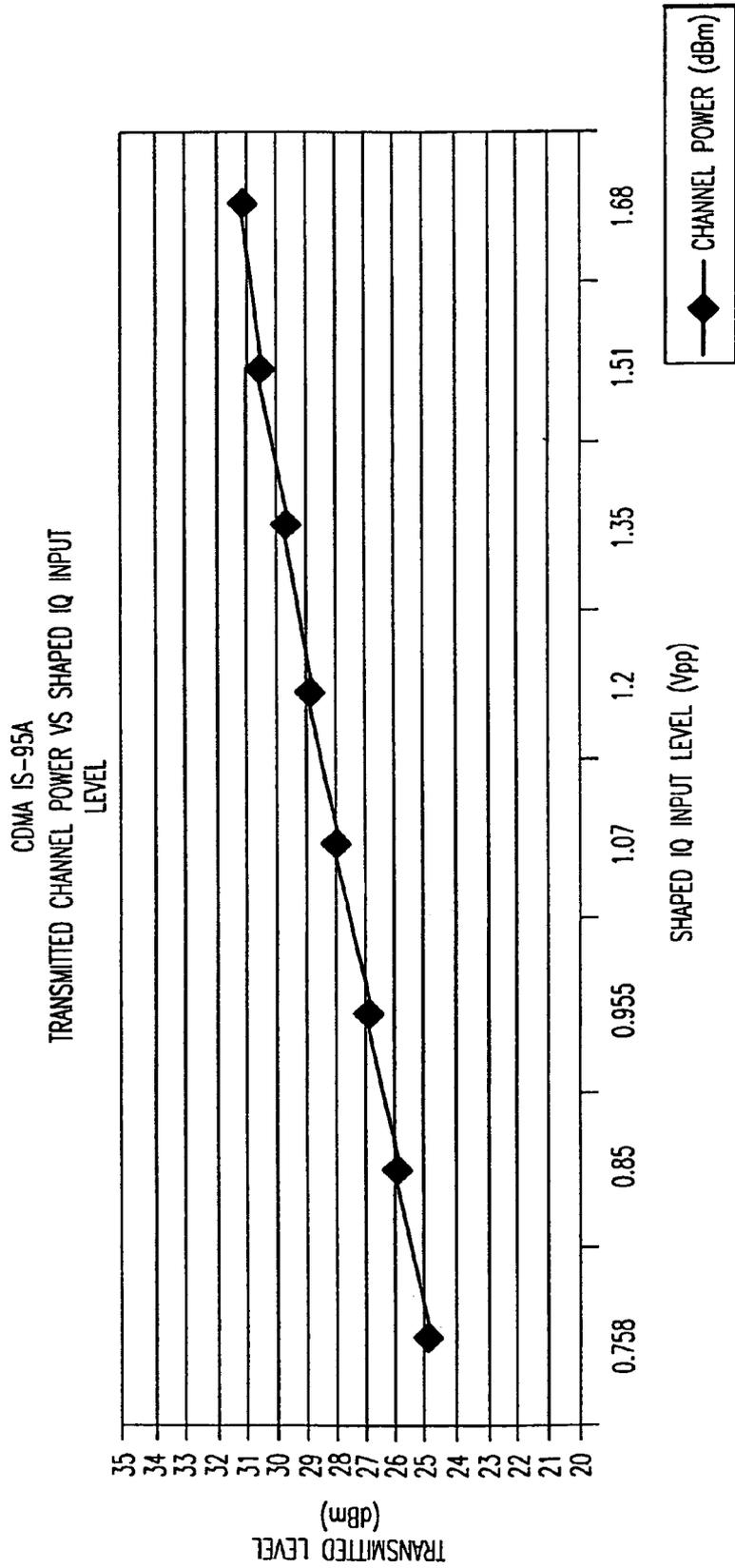


FIG.52C

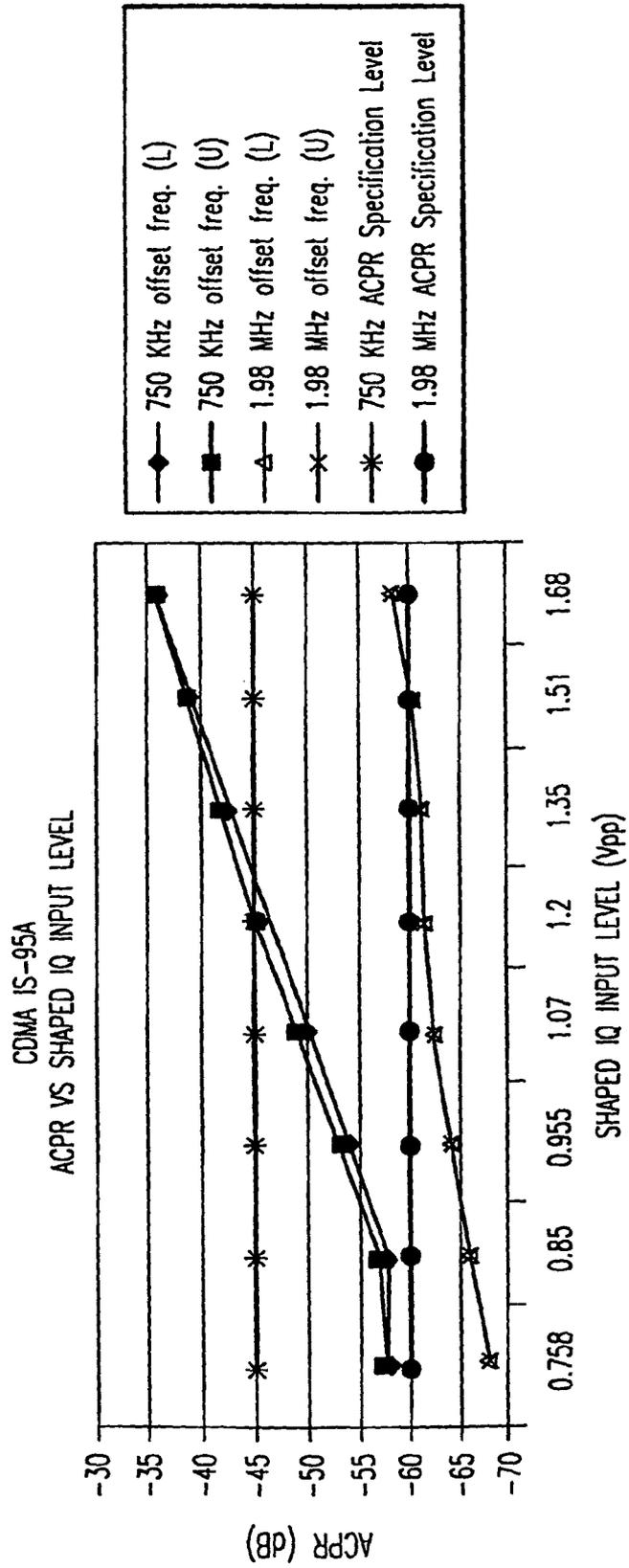


FIG.52D

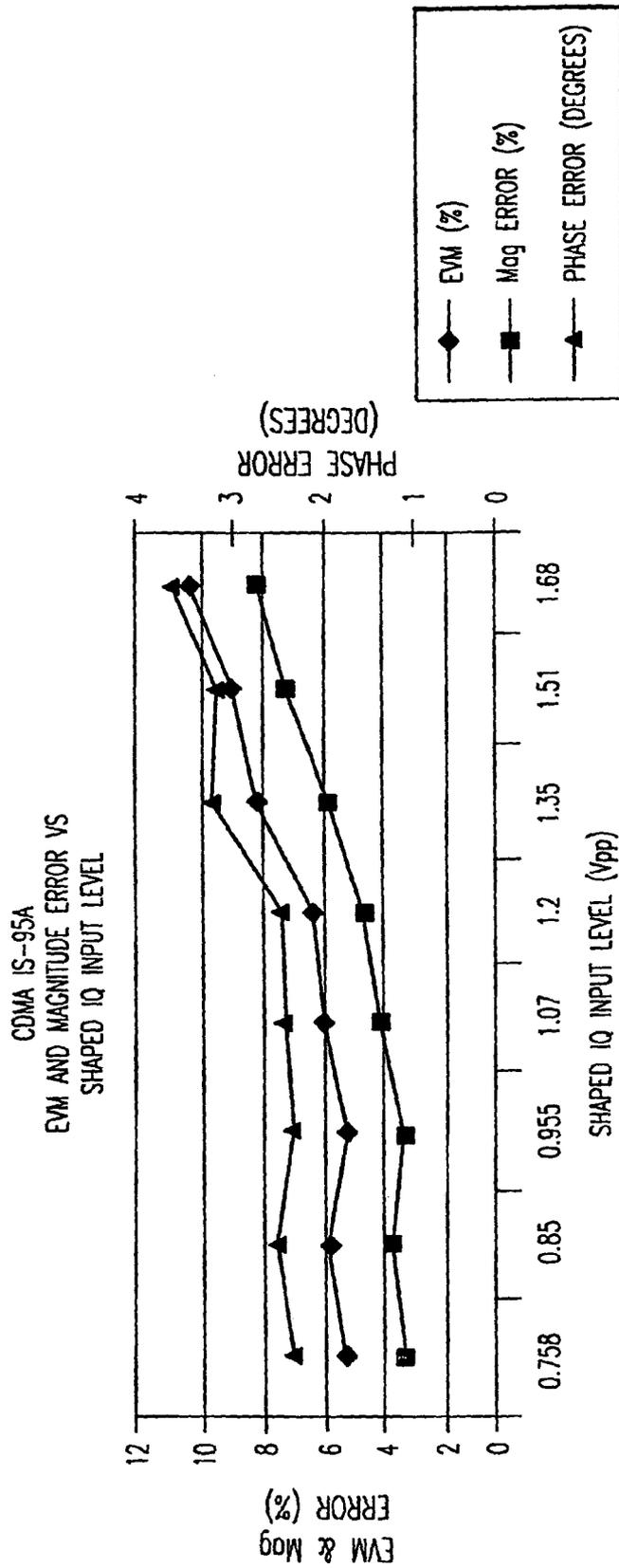


FIG.52E

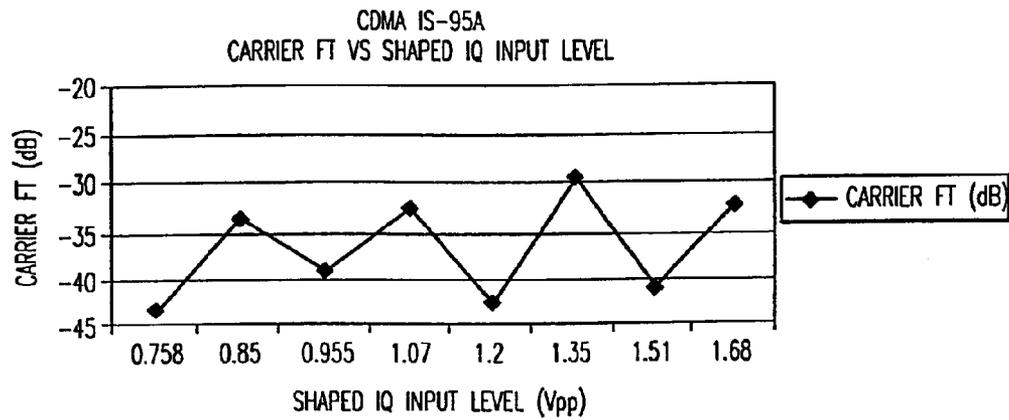


FIG.52F

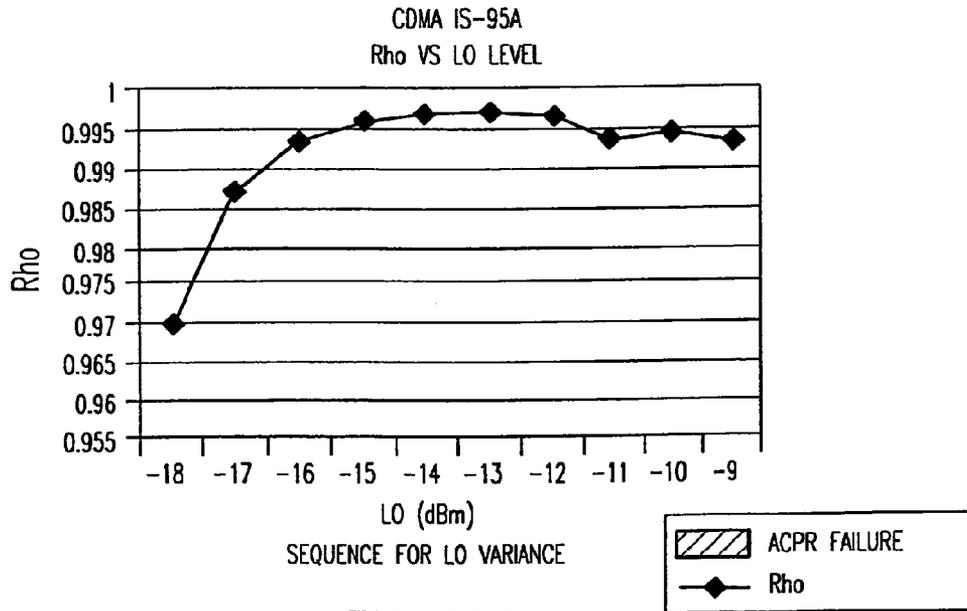


FIG.52G

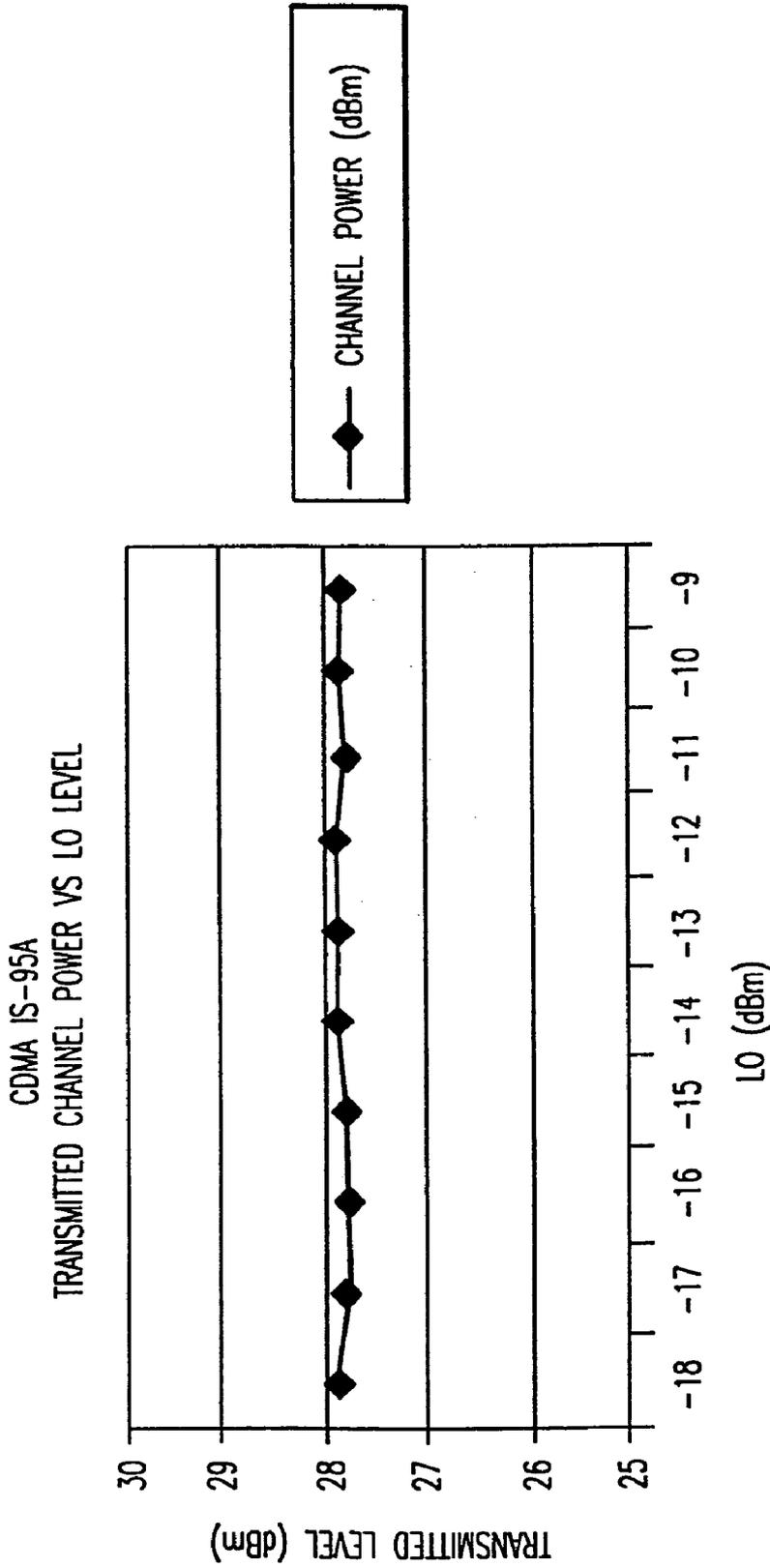


FIG.52H

CDMA IS-95A  
ACPR vs LO LEVEL

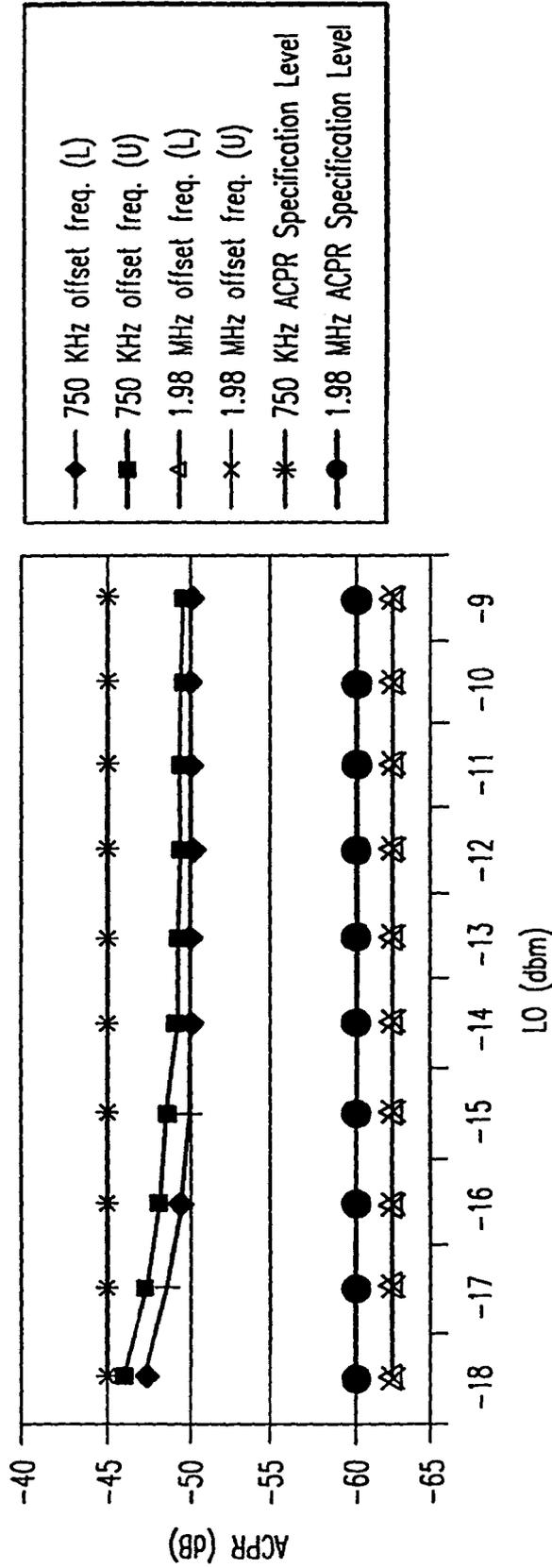
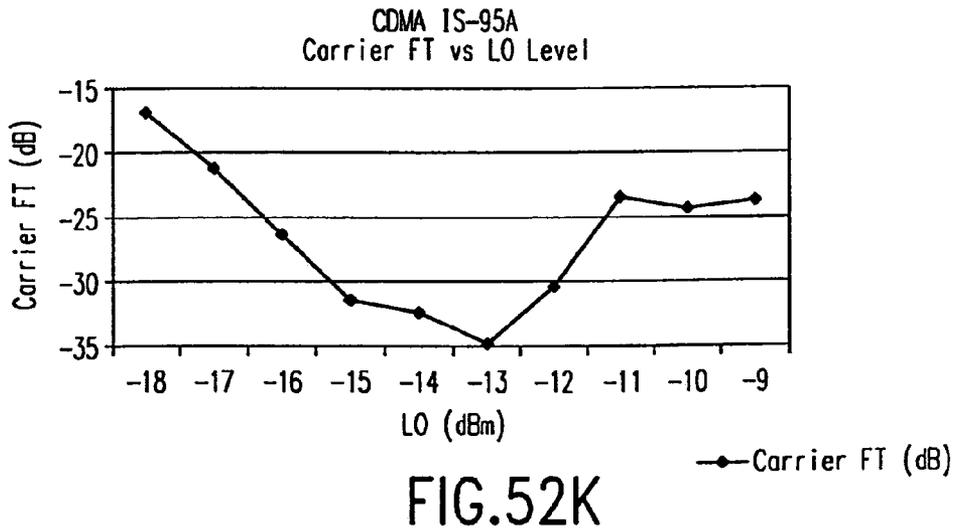
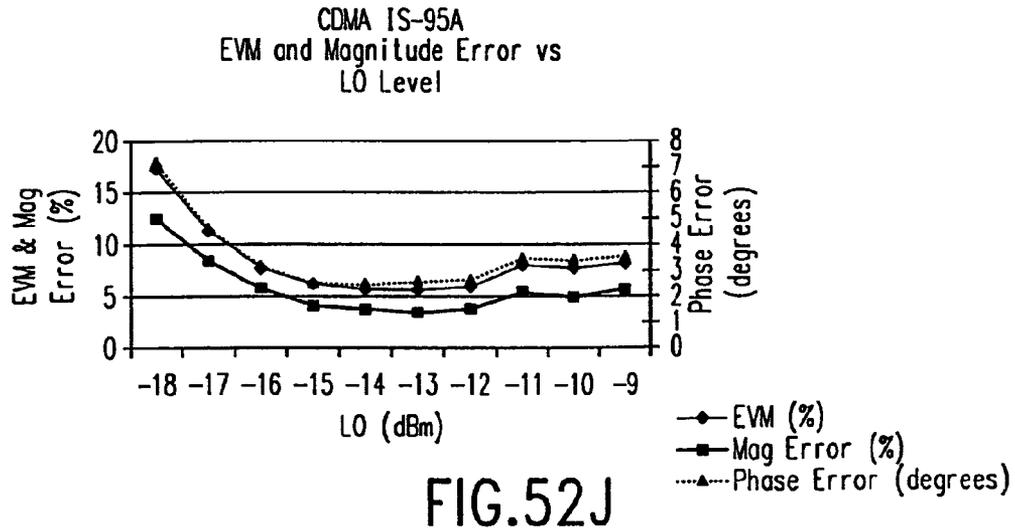


FIG. 521



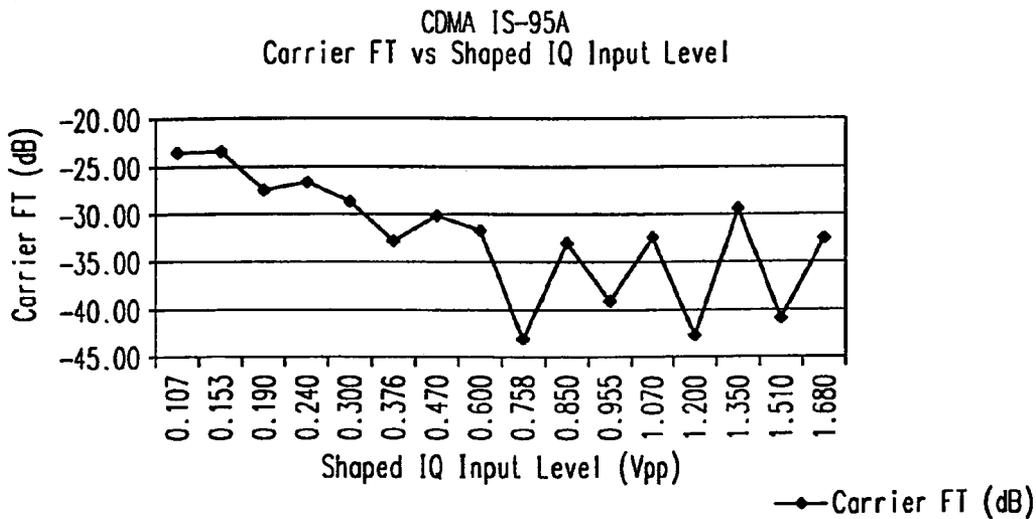
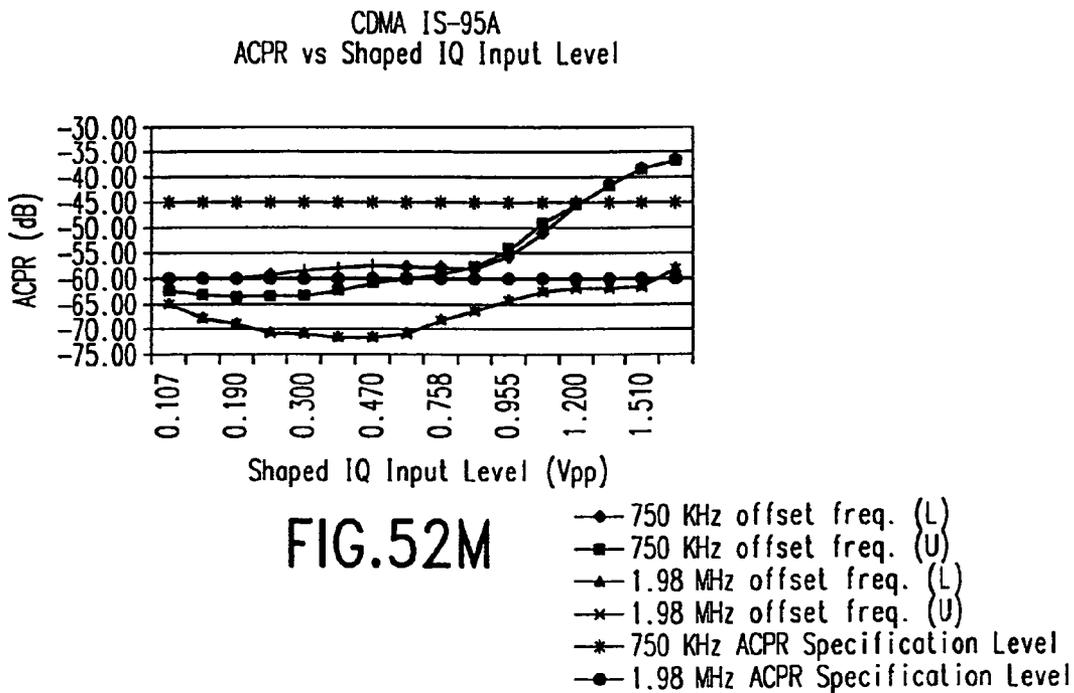


FIG.52L



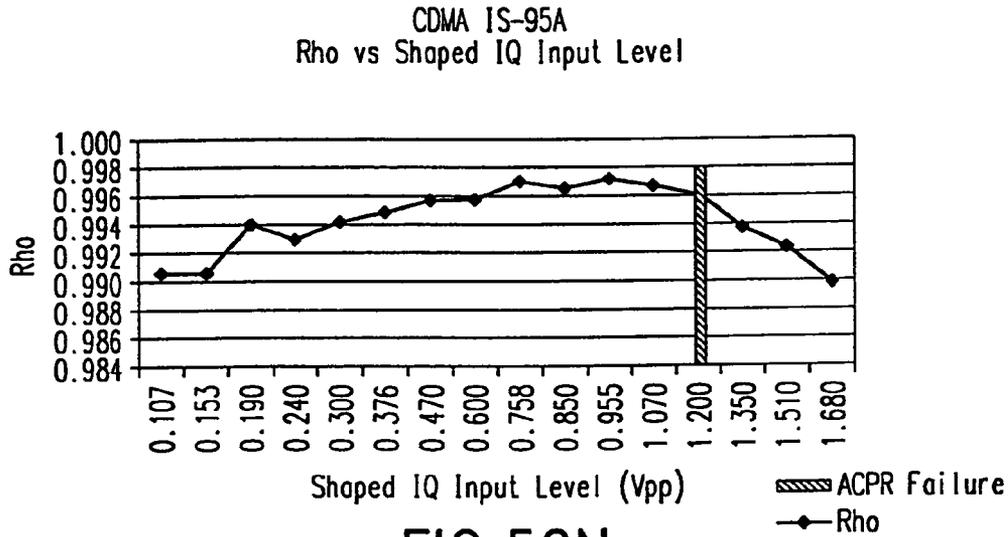


FIG.52N

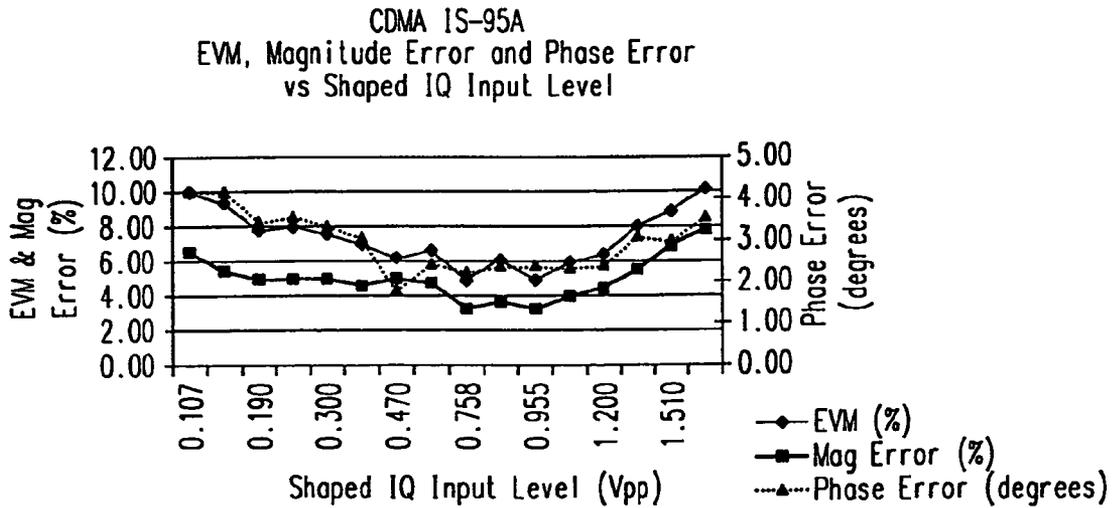


FIG.52O

Sequence For IQ Input Level Variance  
CDMA IS-95A Mobile Transmitter@+3.3V  
Rho vs Shaped IQ Input Level

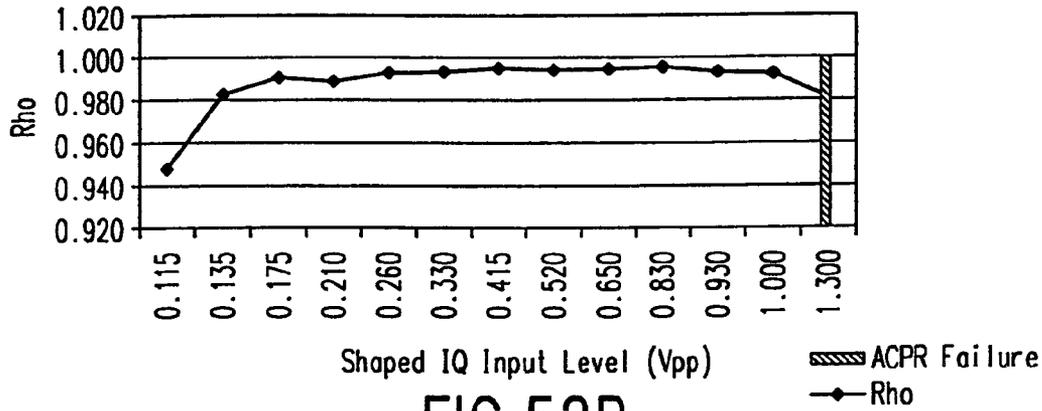


FIG.52P

CDMA IS-95A Mobile Transmitter@+3.3V  
Transmitted Channel Power vs Shaped IQ Input Level

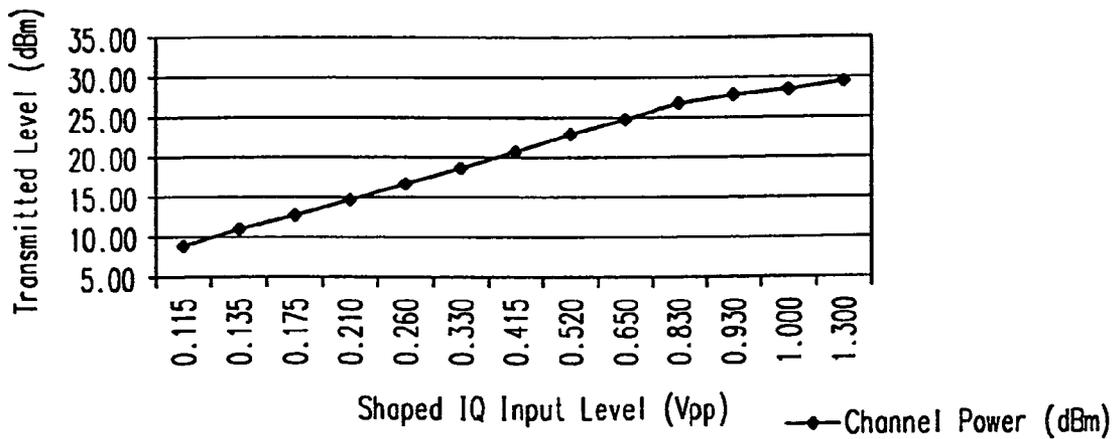
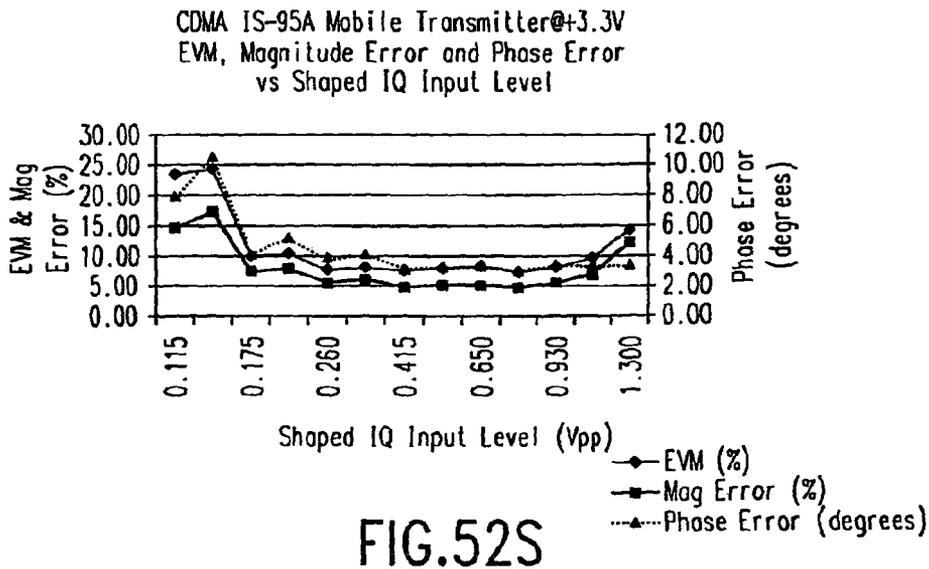
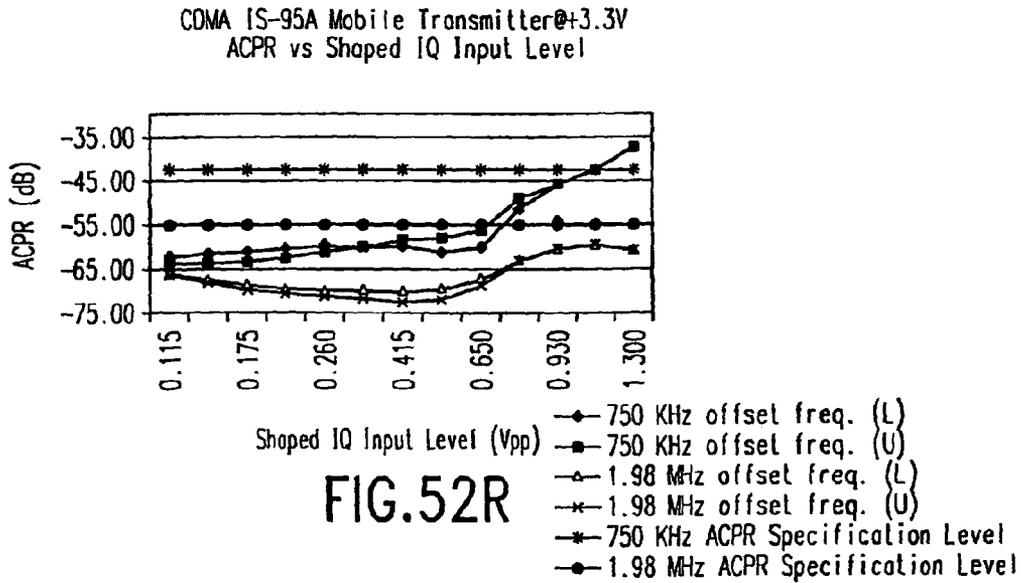


FIG.52Q



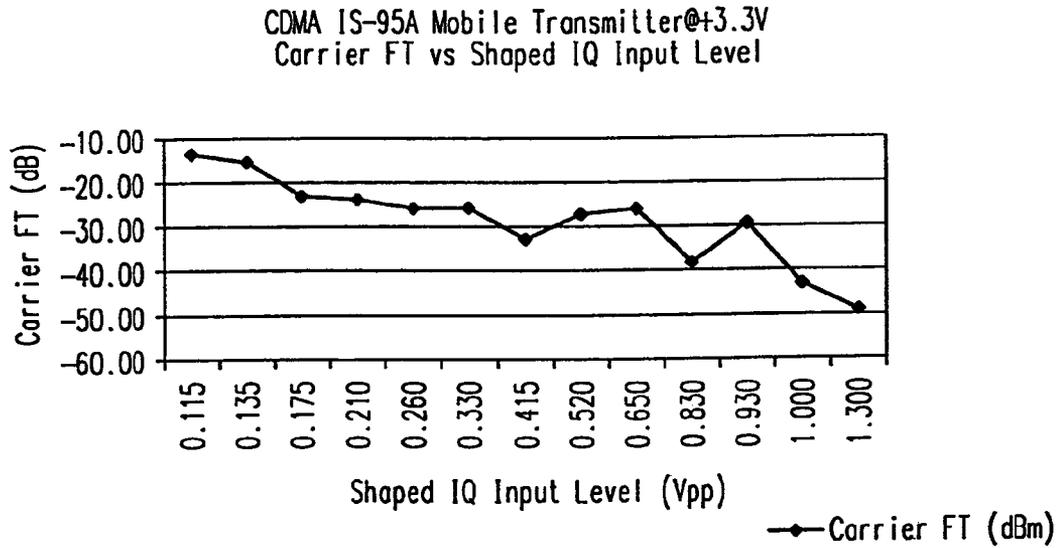


FIG.52T

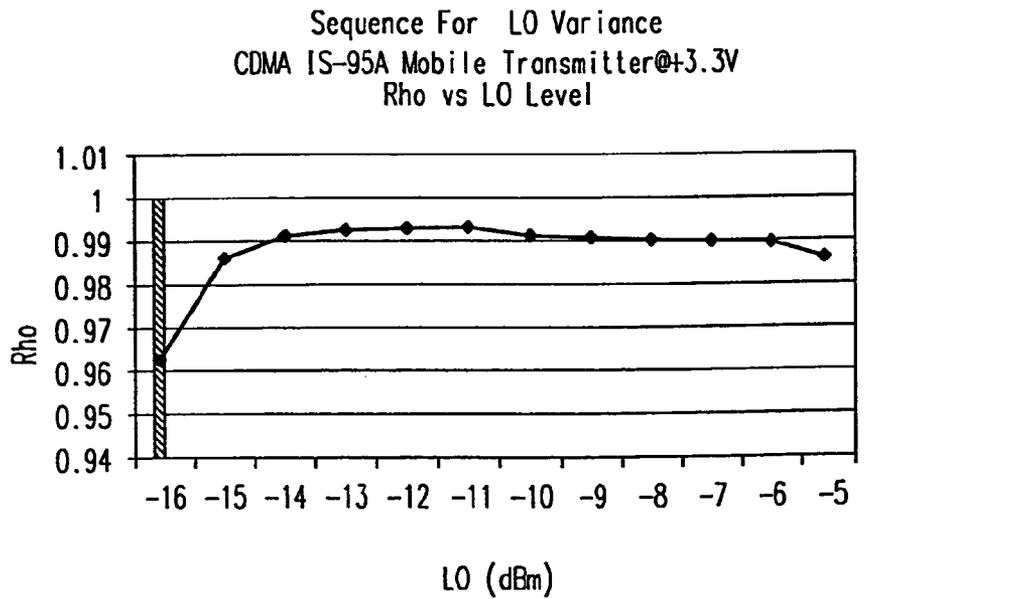


FIG.52U

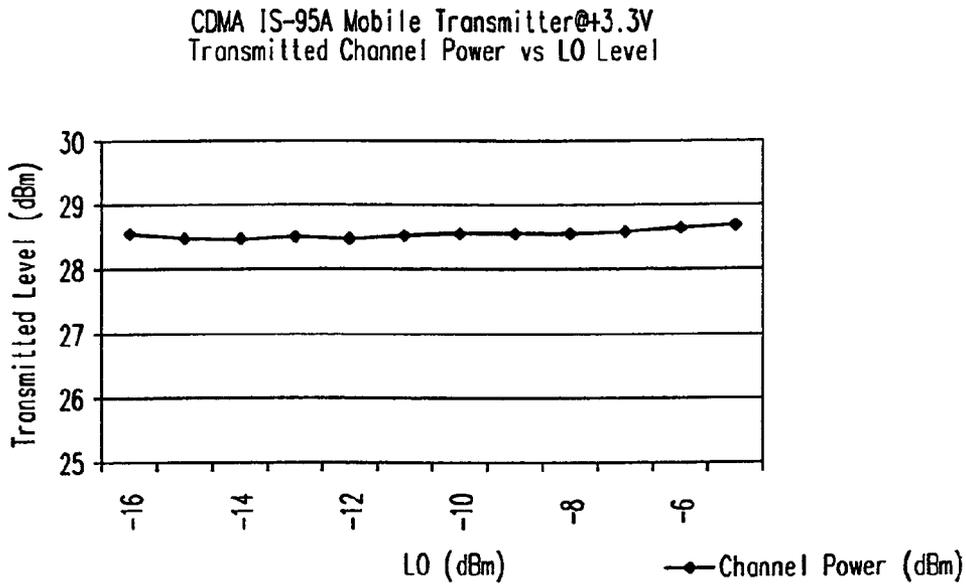


FIG.52V

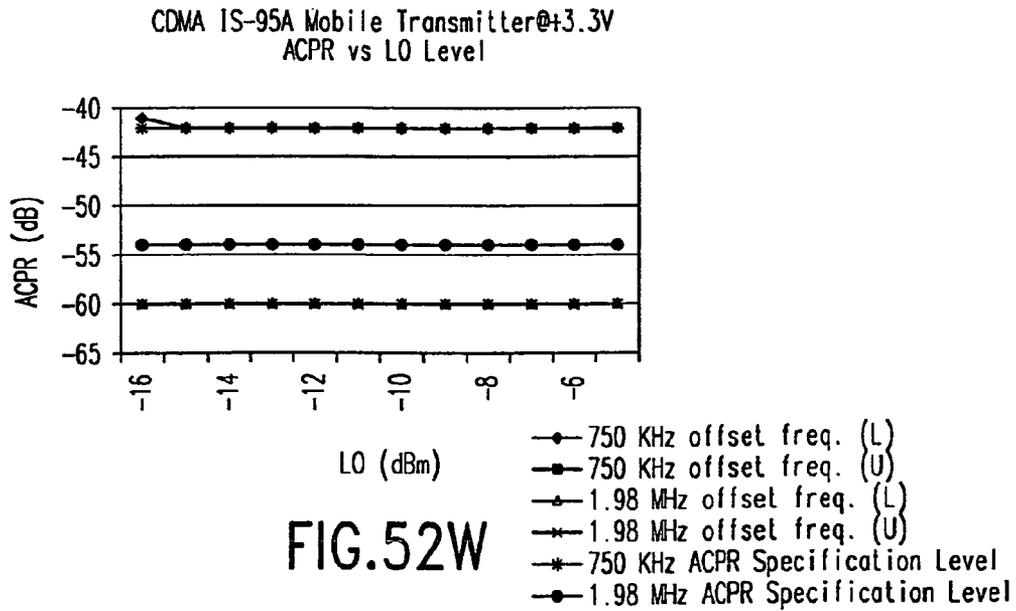
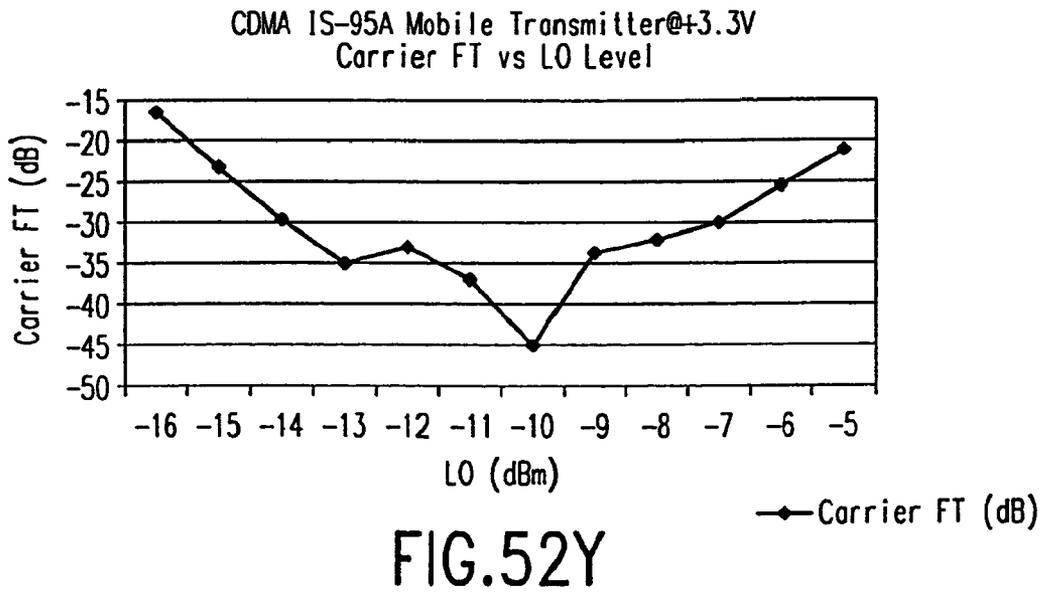
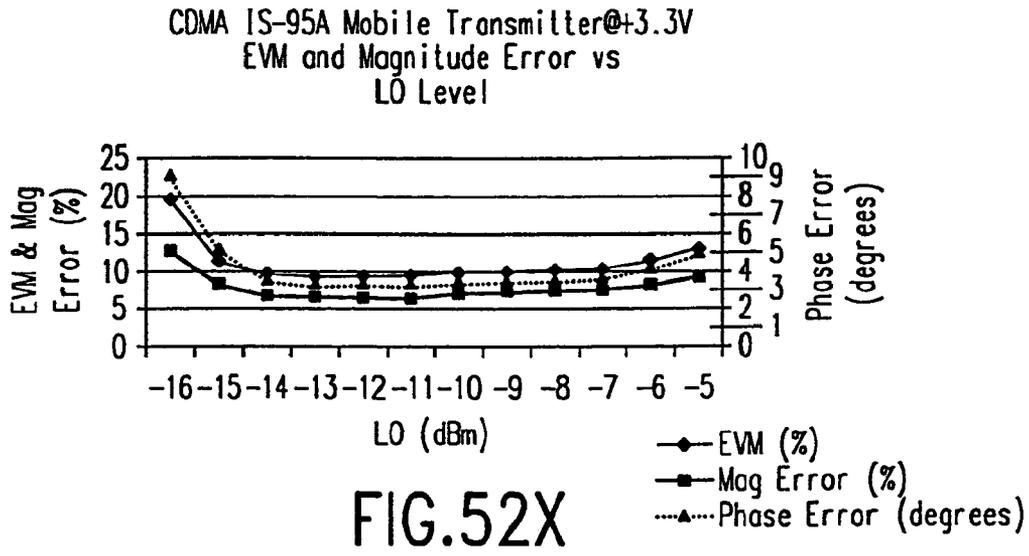


FIG.52W



QUANTITY	DESCRIPTION	VOLTAGE	TOTAL CURRENT	POWER
2	CORES	3.3	4mA	13.2mW
2	BASEBAND INTERFACE CIRCUITS WITH/BW LIMIT	3.3	6mA	21.8mW
1	CLOCK CIRCUIT	3.3	5mA	20.0mW
			SUB TOTAL	54.0mW

FIG. 52Z

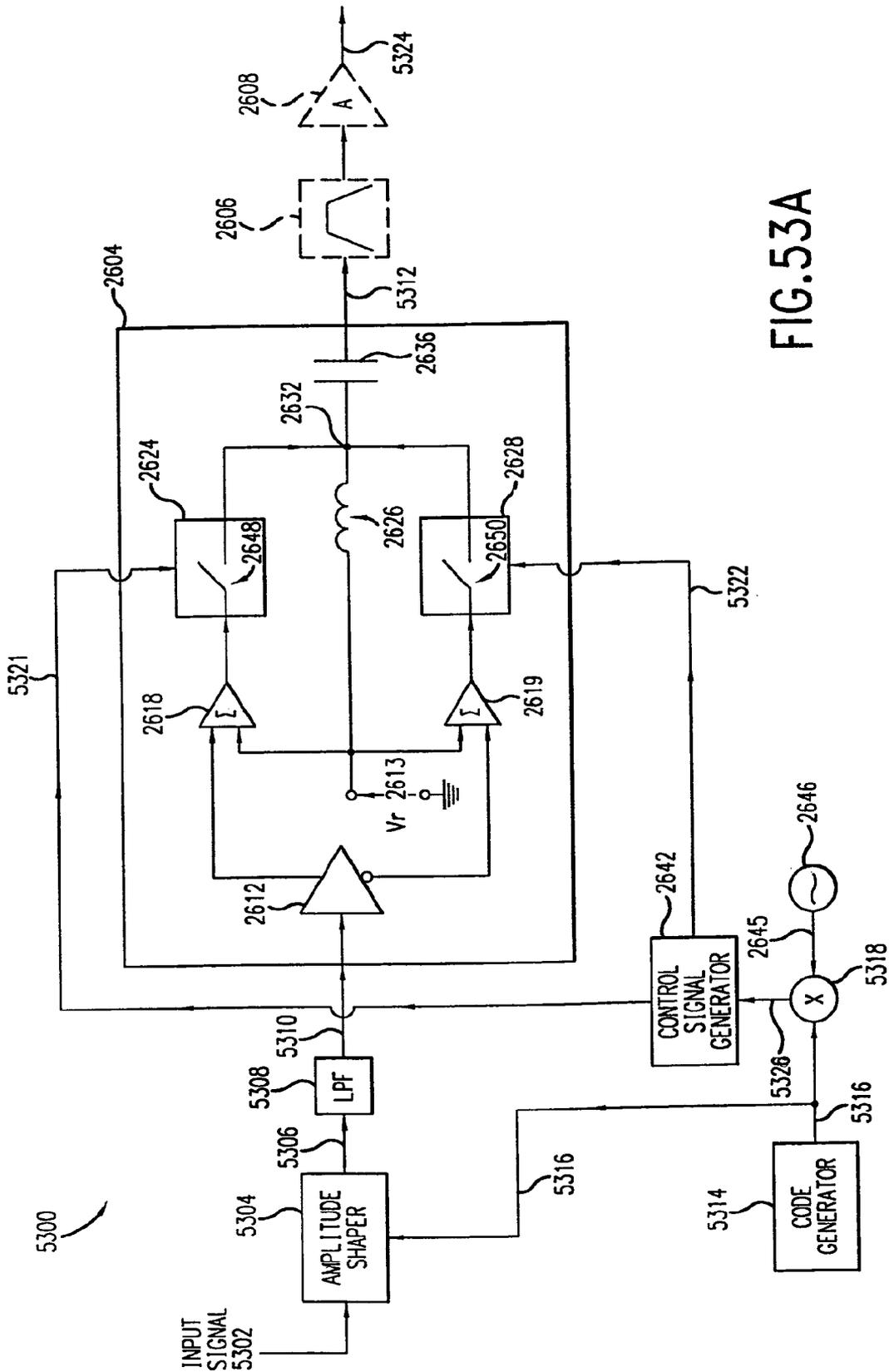


FIG. 530A

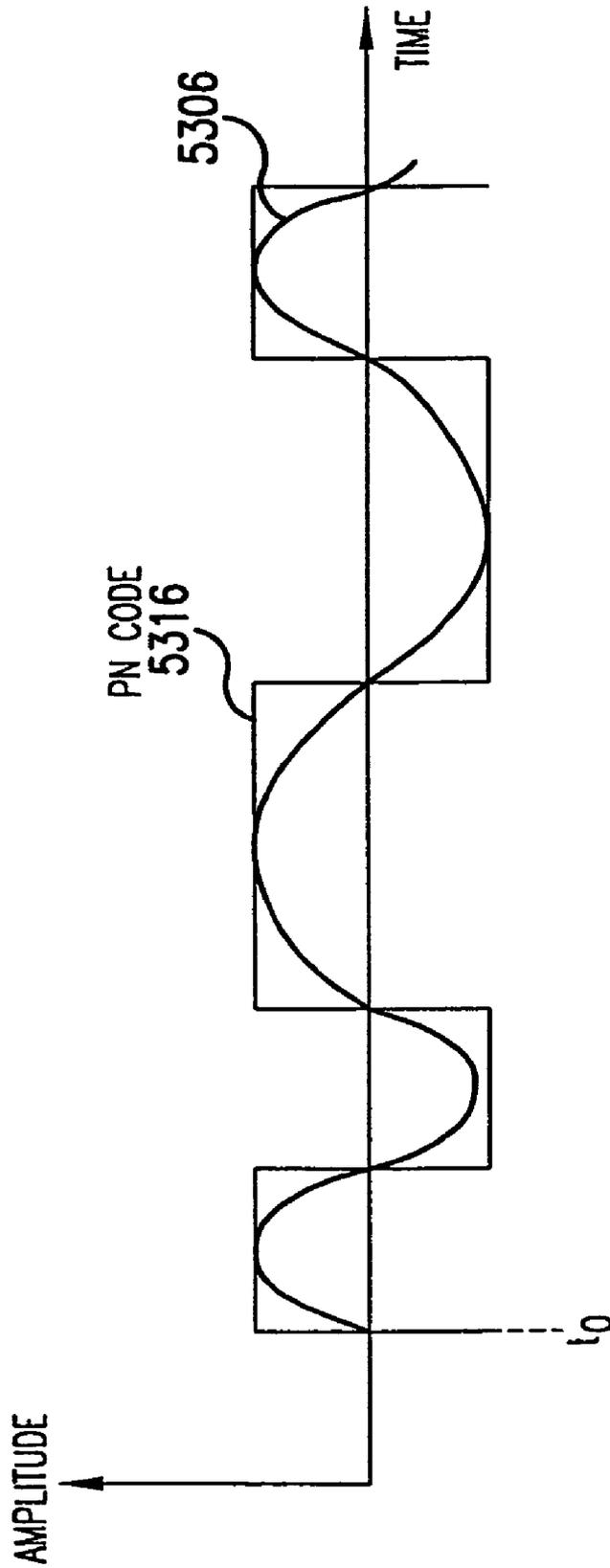


FIG. 53B

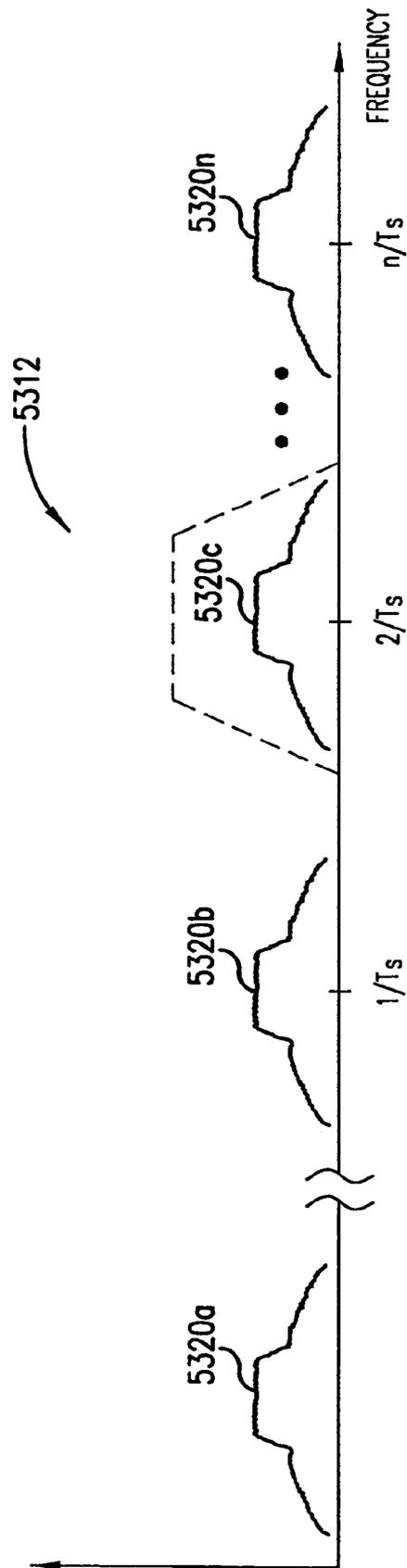


FIG. 53C

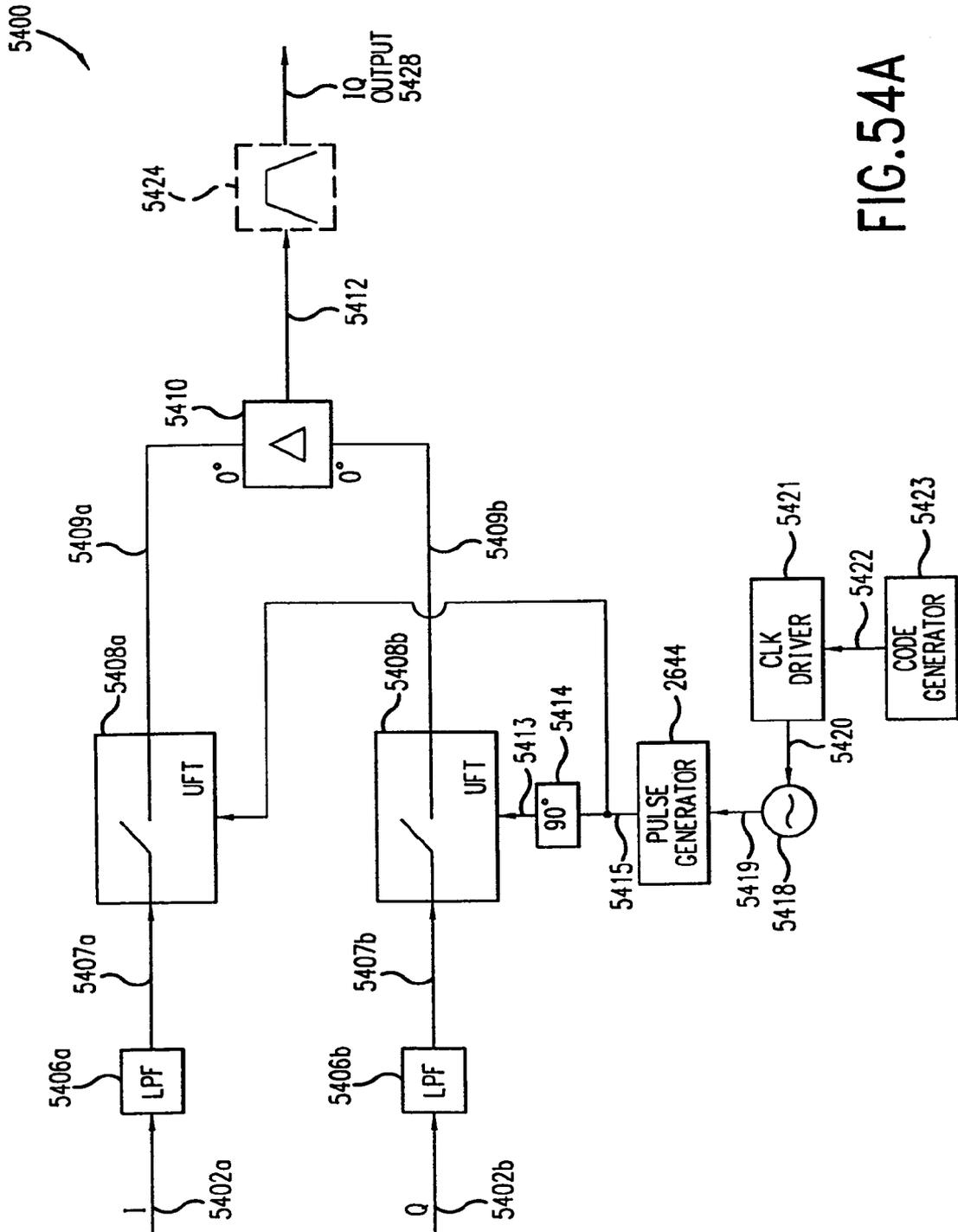


FIG. 54A

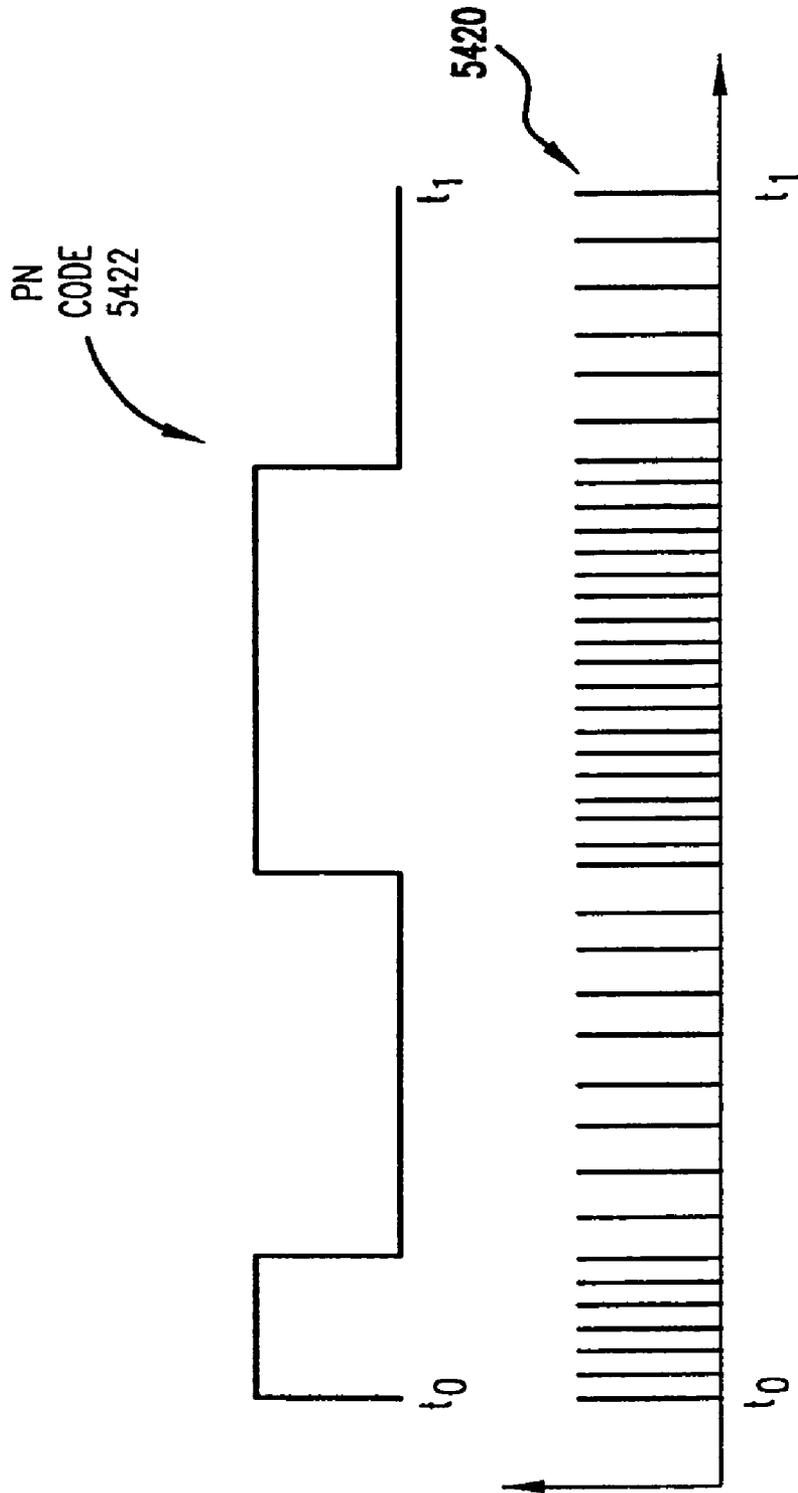


FIG. 54B

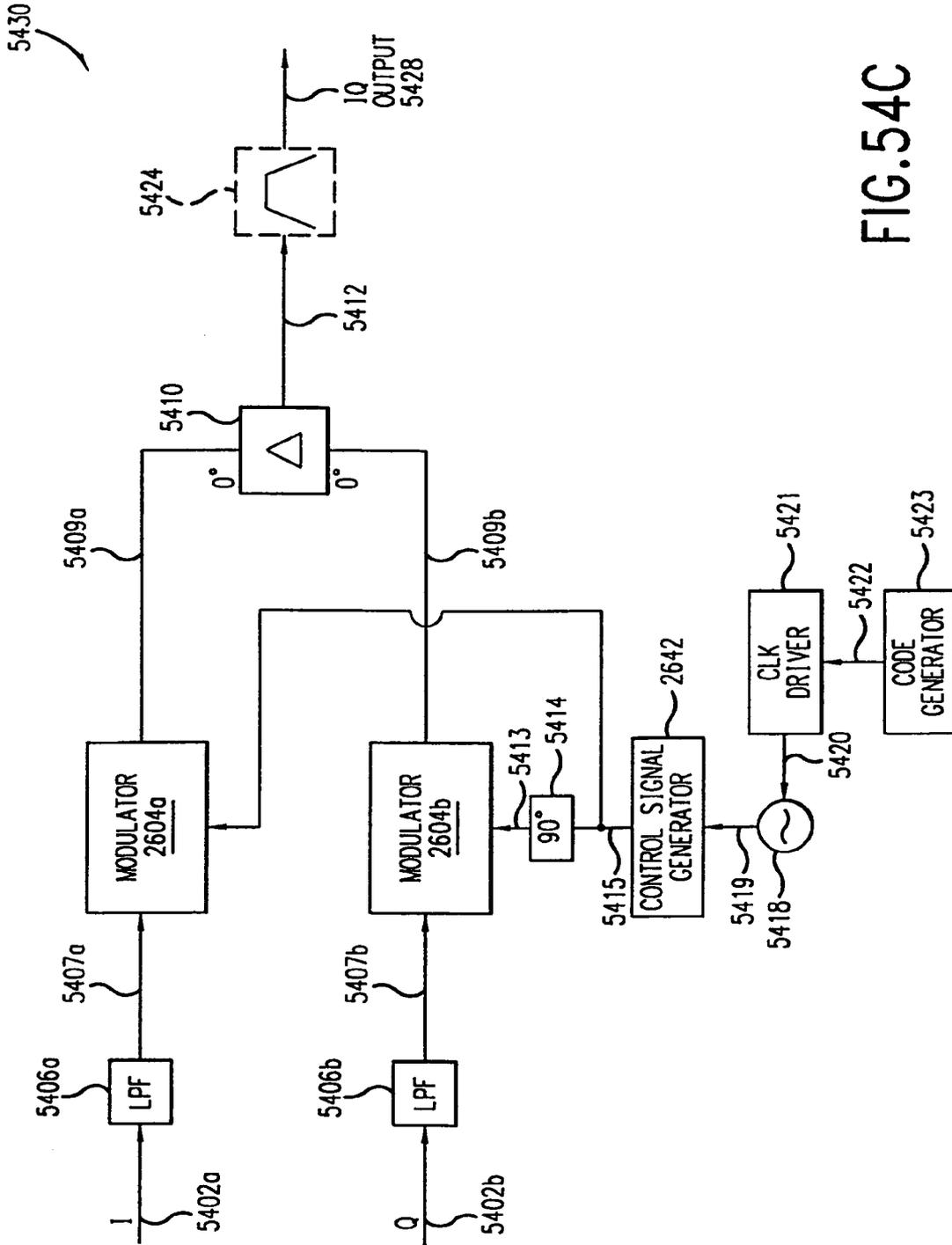


FIG. 54C

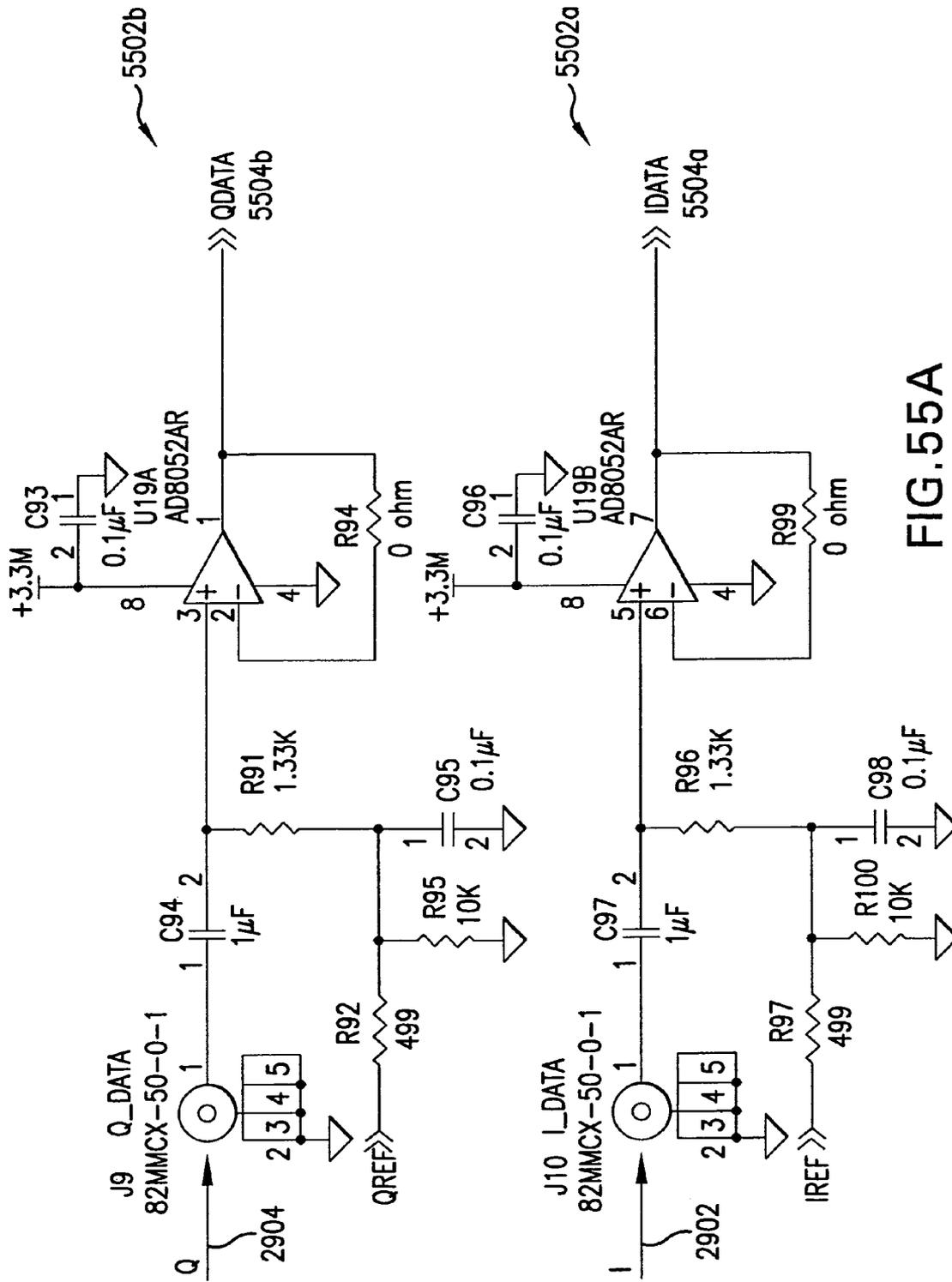


FIG. 55A

FIG.55 B-1	FIG.55 B-2	FIG.55 B-3	FIG.55 B-4
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FIG.55B

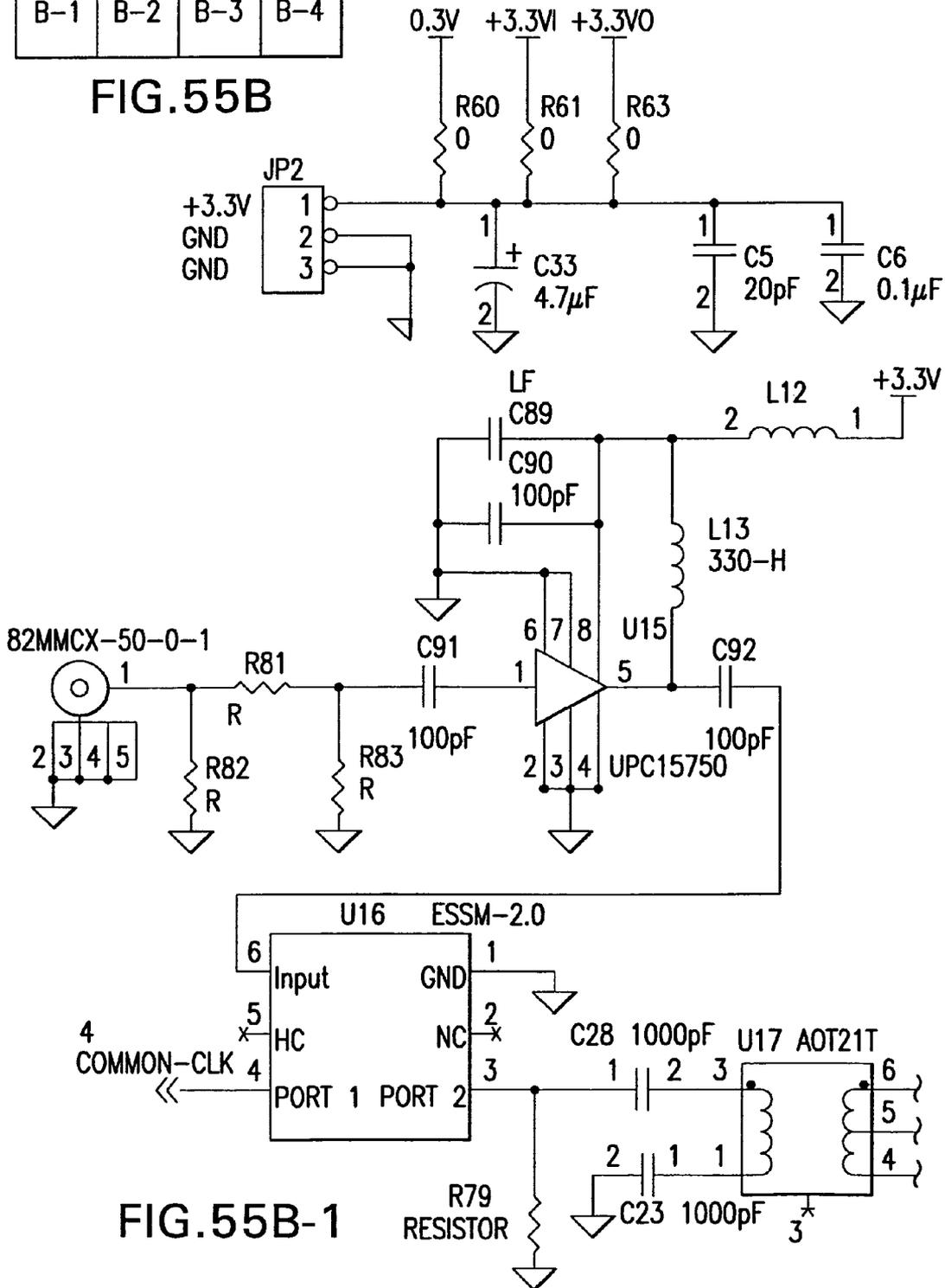


FIG.55B-1

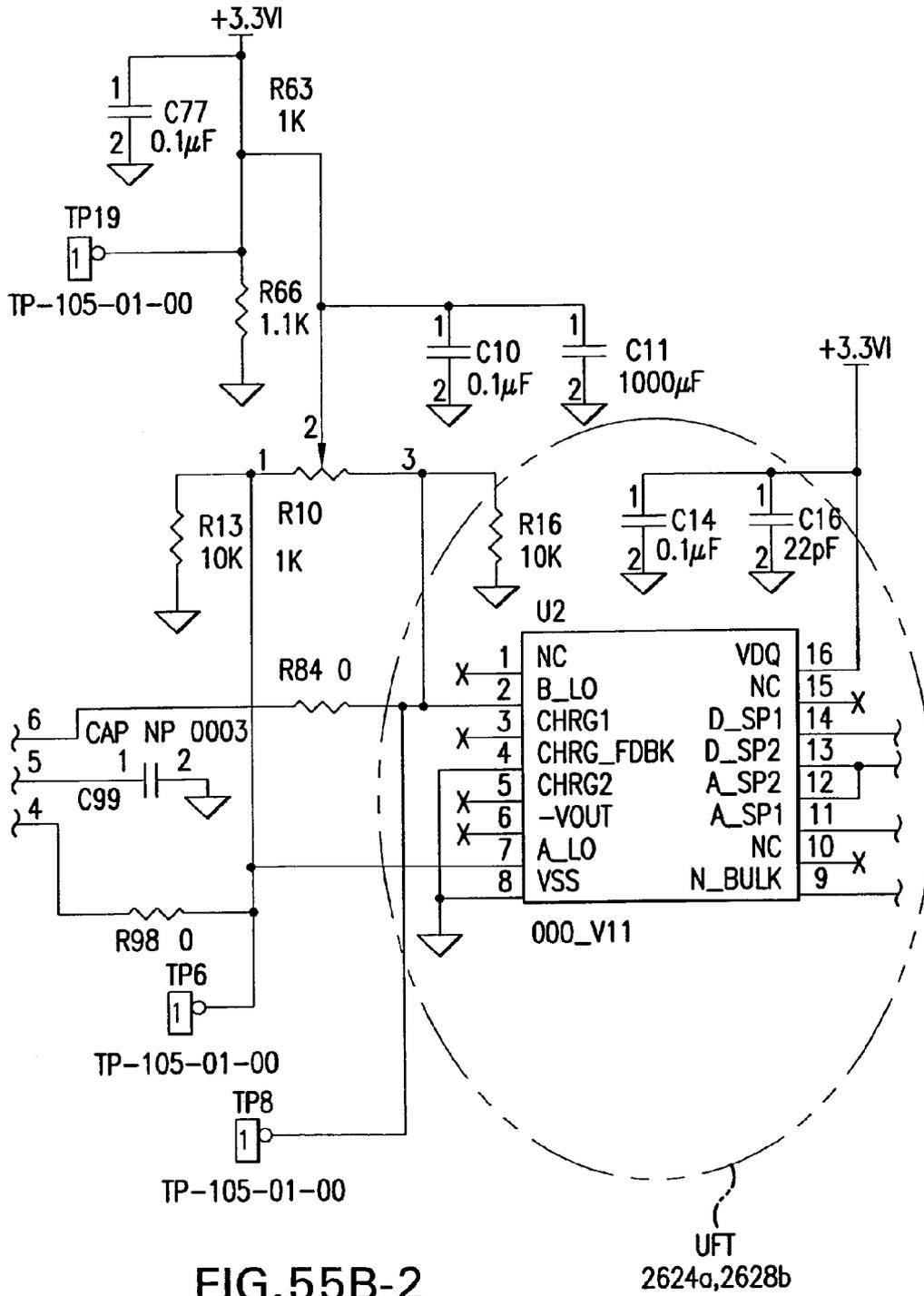


FIG.55B-2

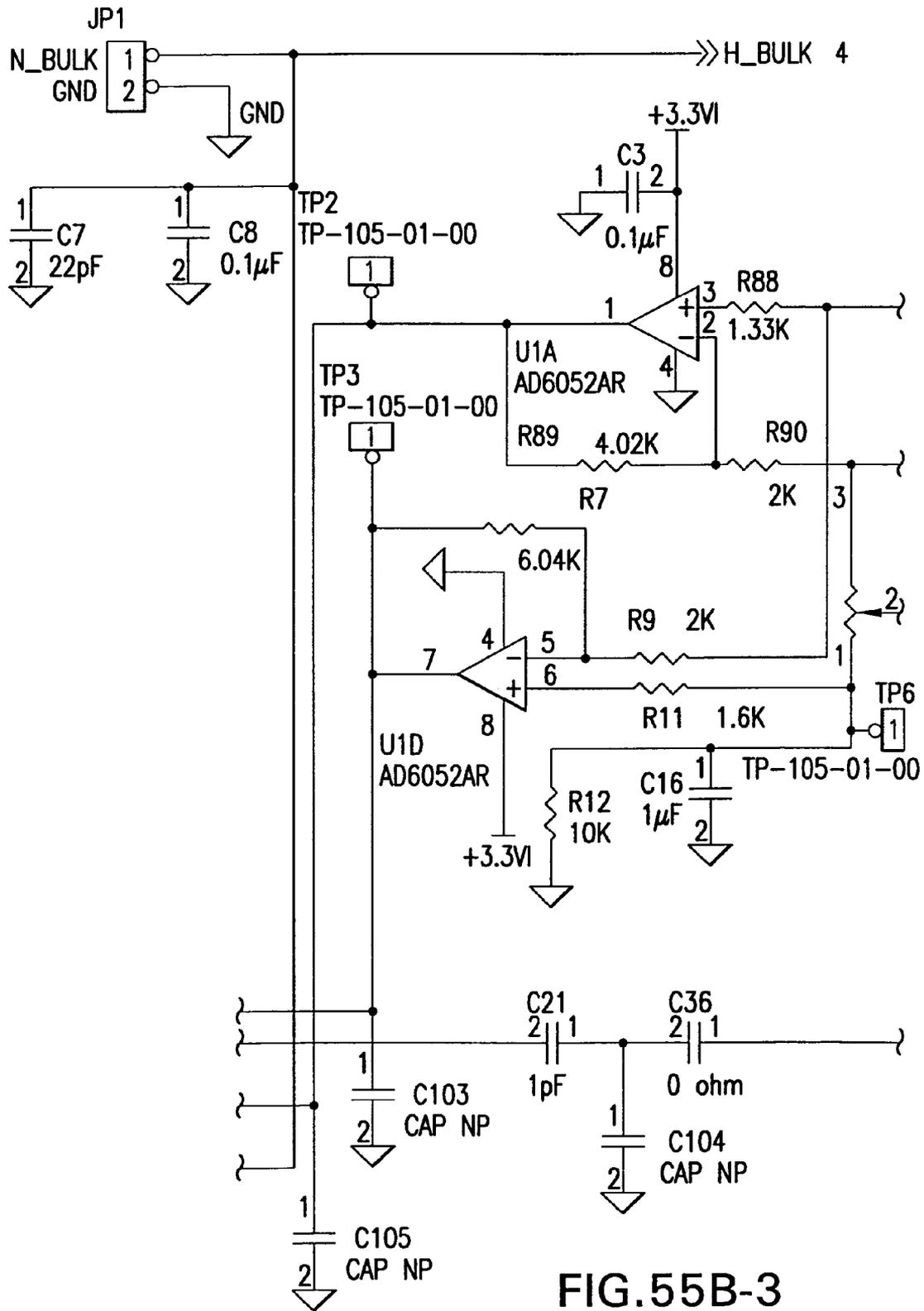


FIG.55B-3

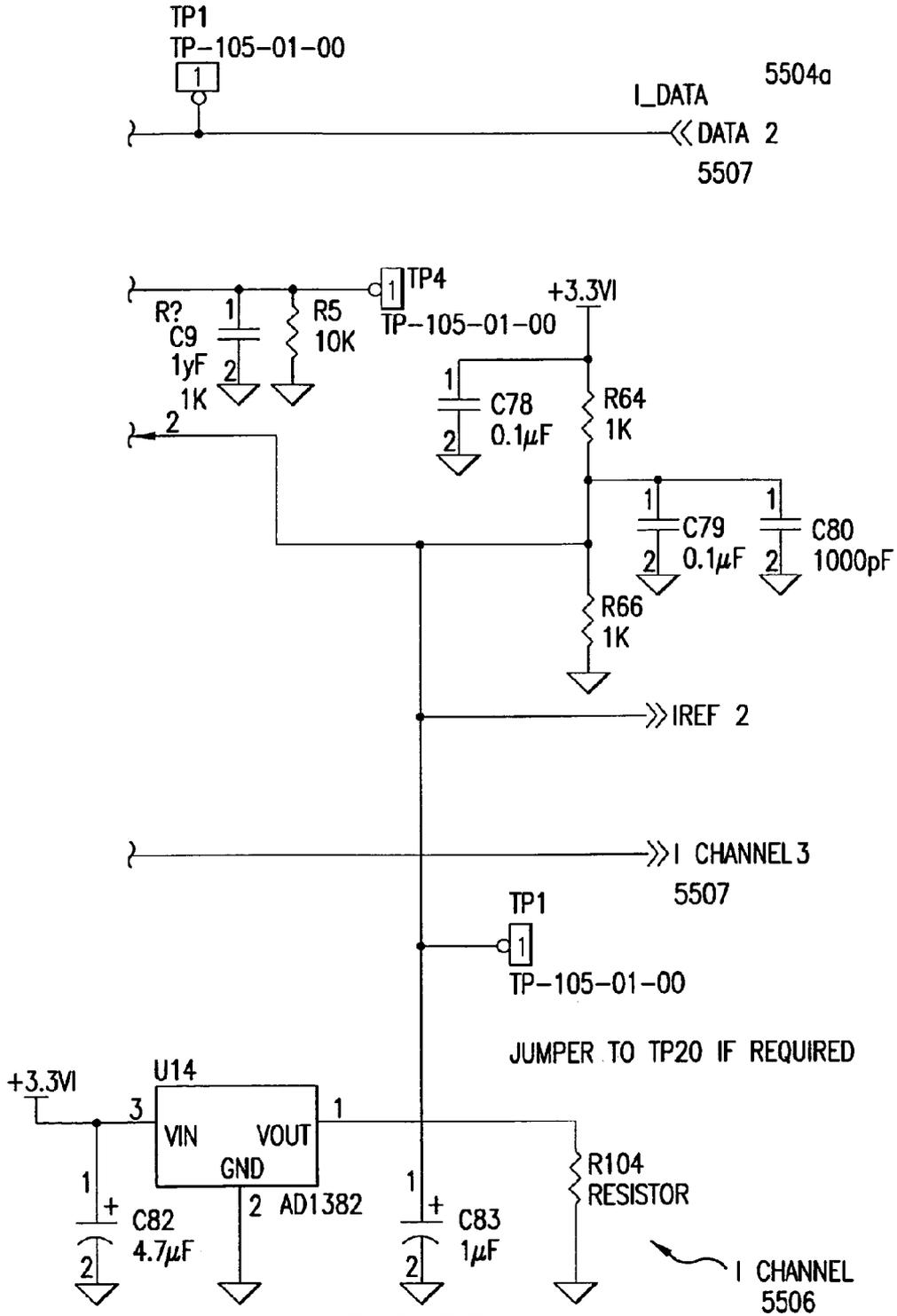


FIG. 55B-4

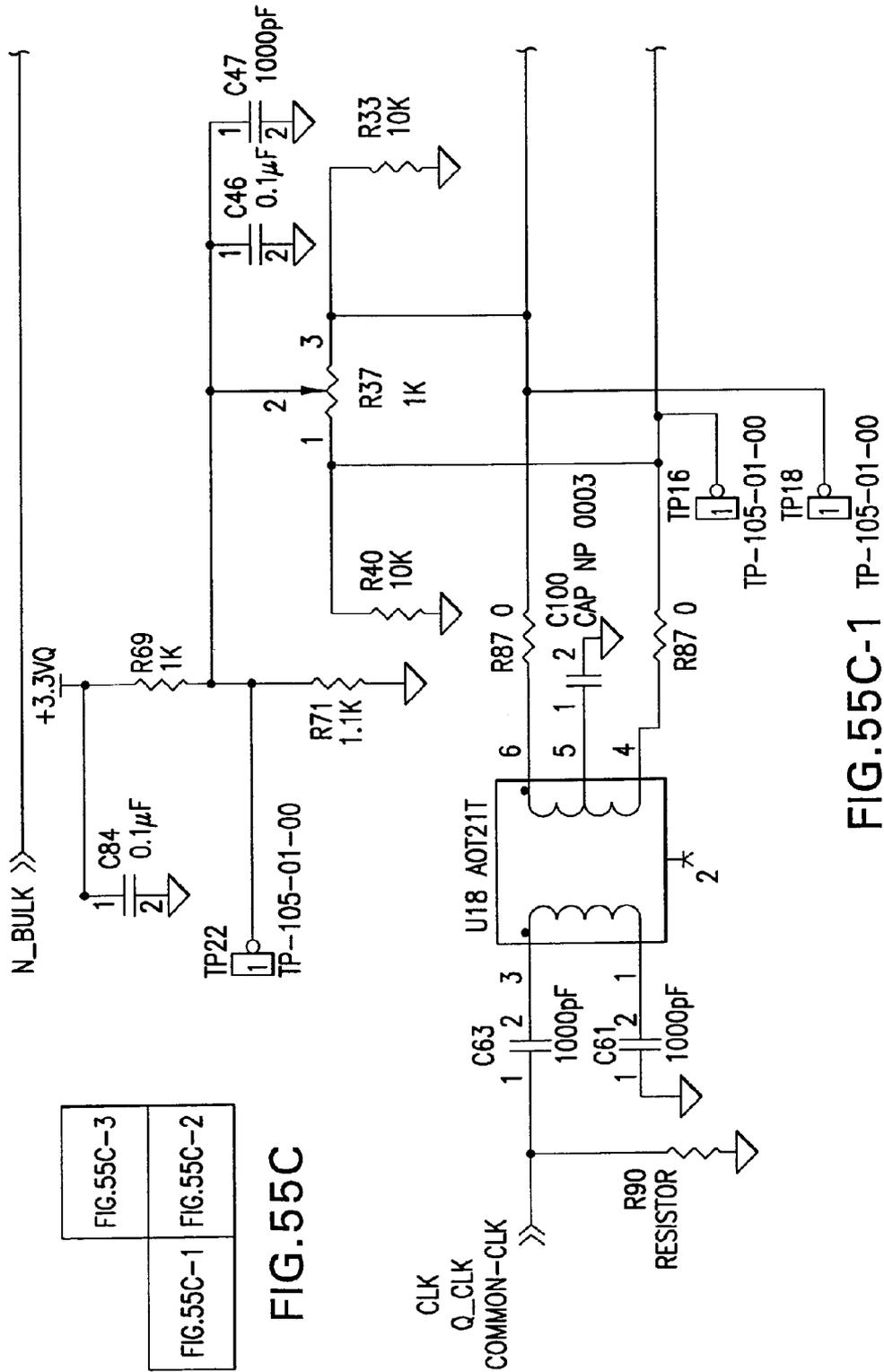


FIG. 555C-1  
FIG. 555C-2  
FIG. 555C-3

FIG. 555C-1 TP-105-01-00



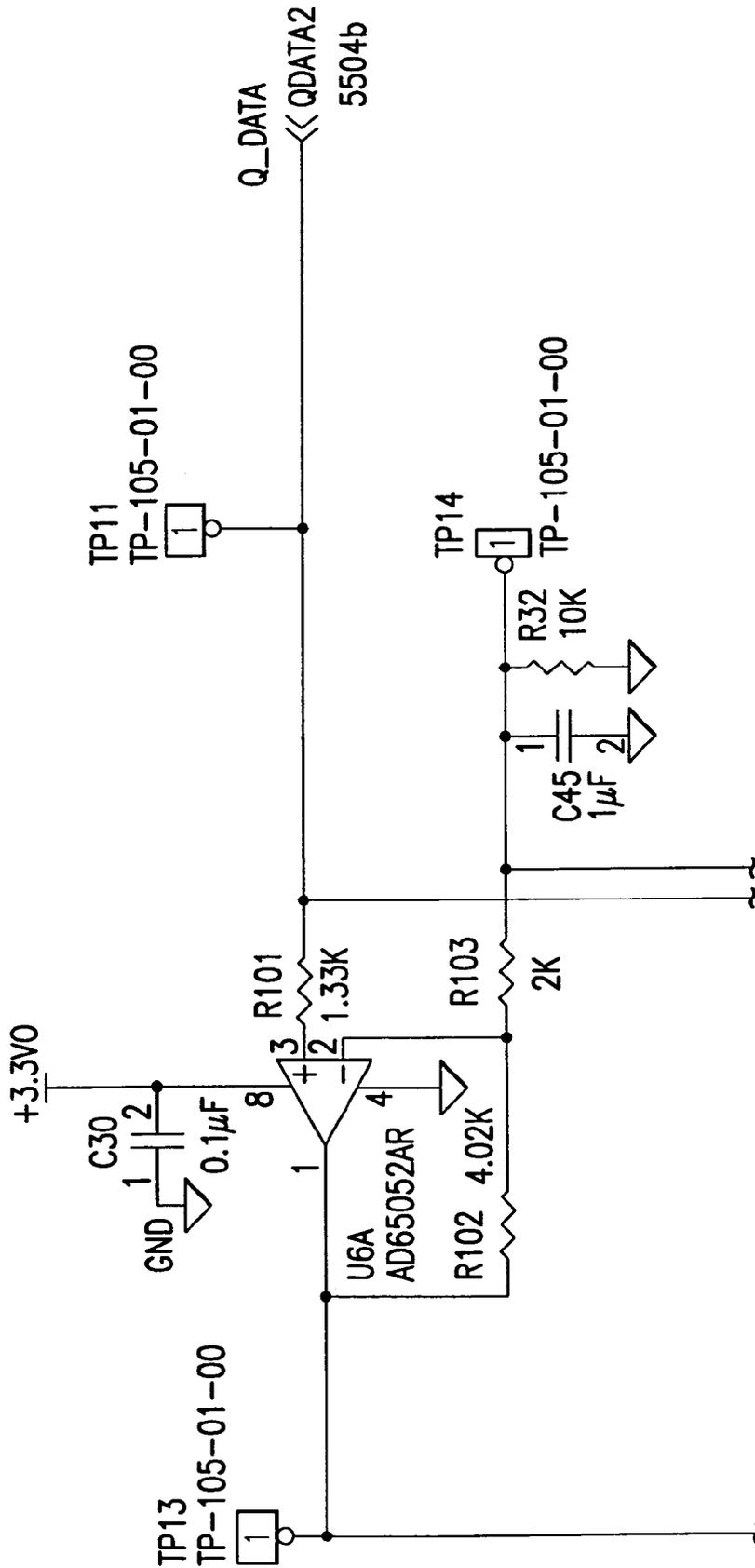


FIG. 55C-3

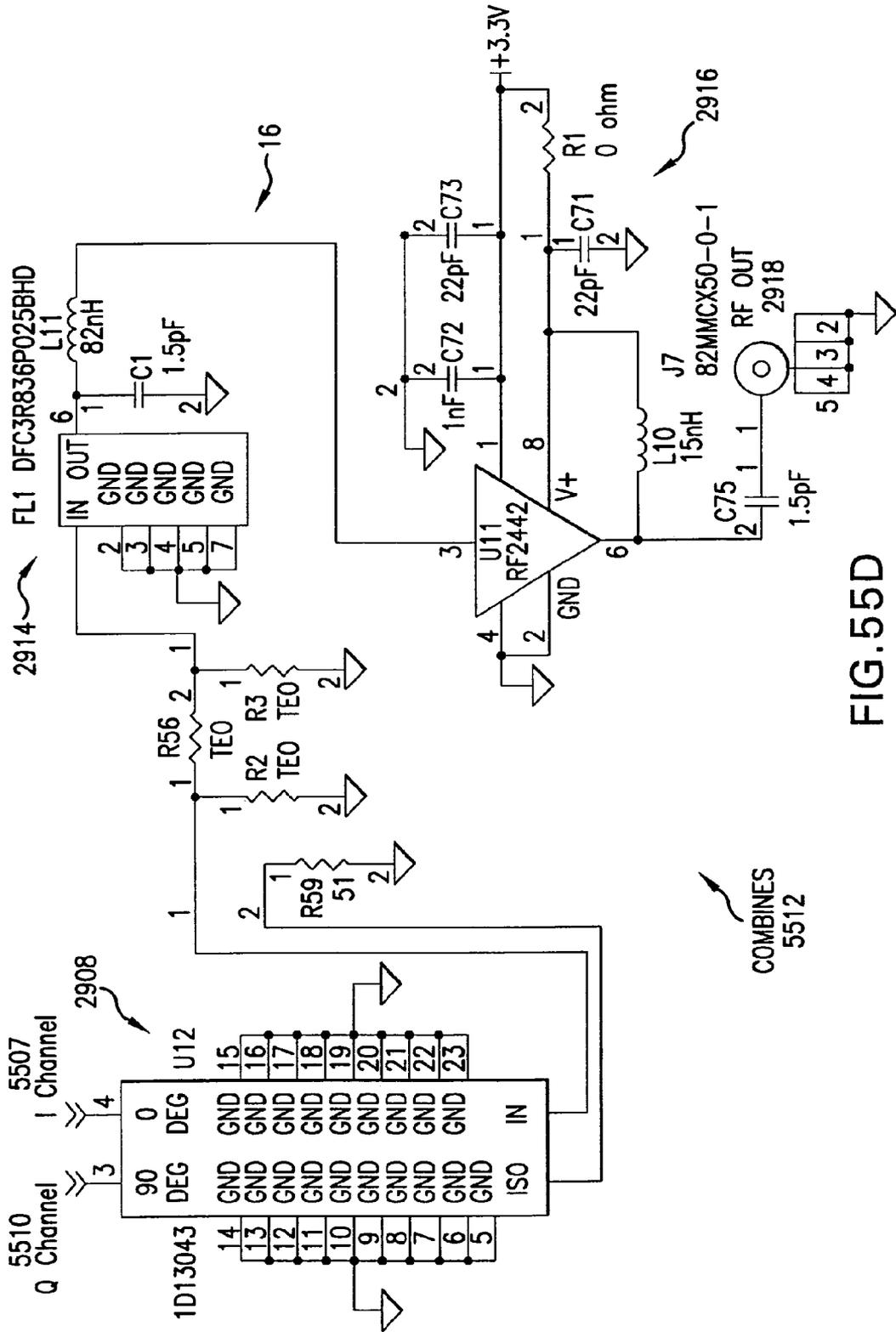


FIG. 555D



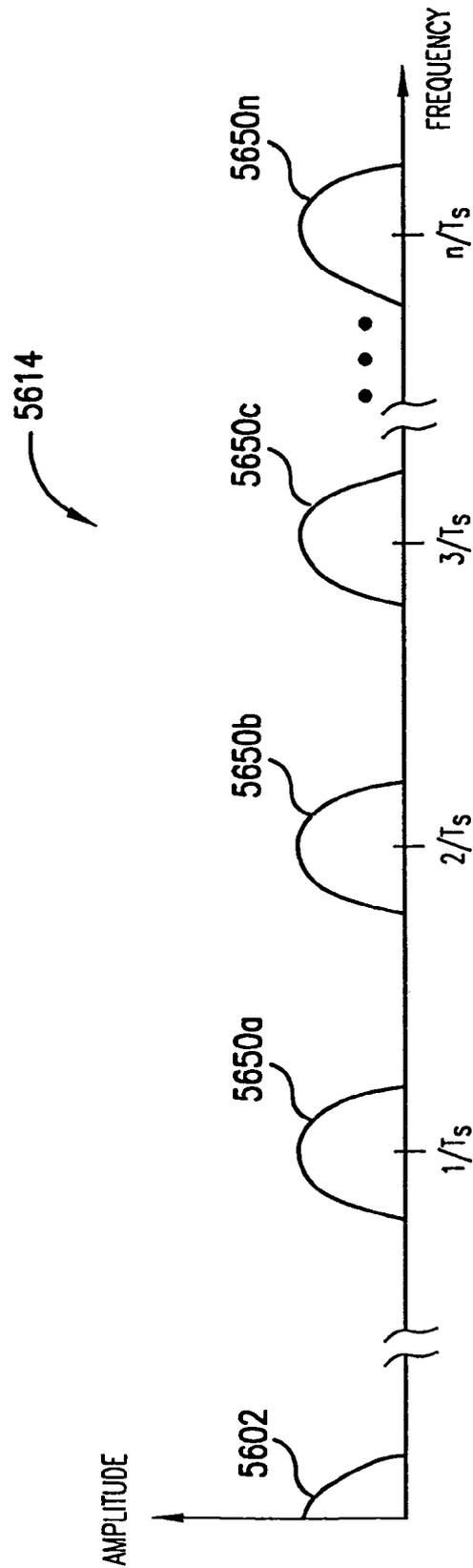


FIG. 56B

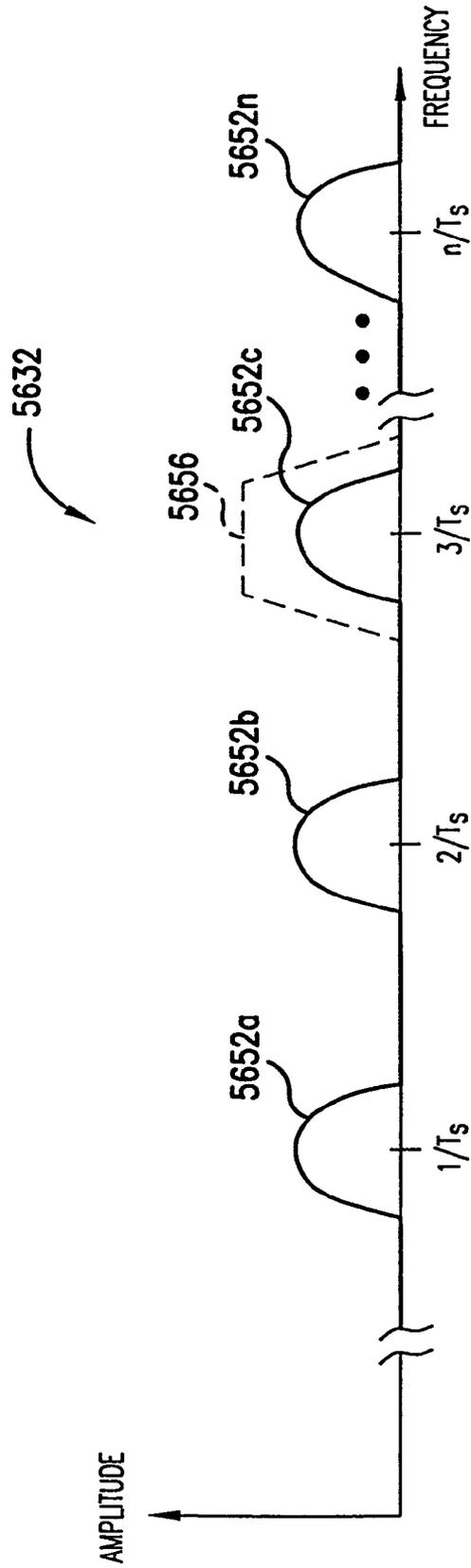


FIG. 56C



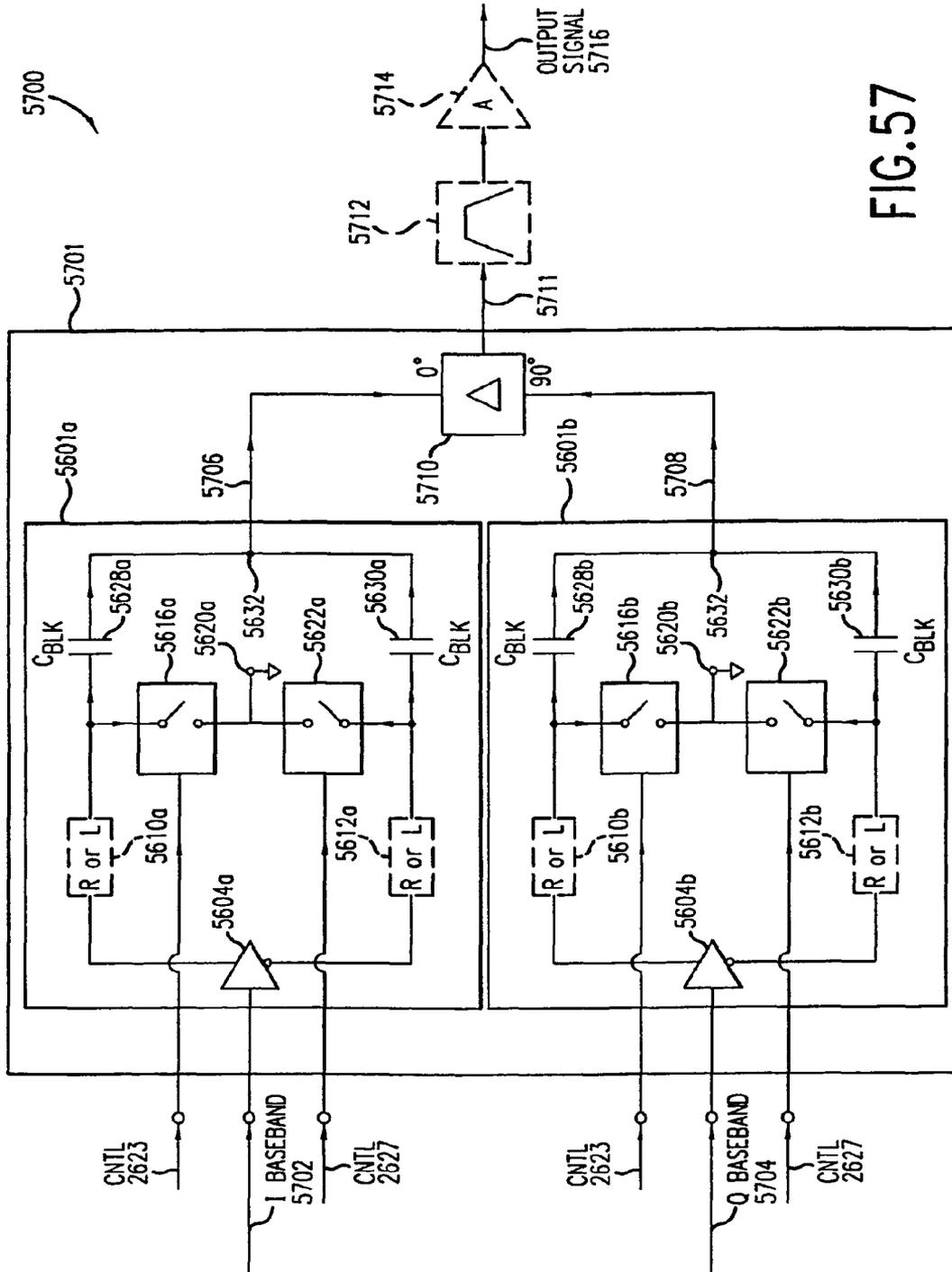


FIG. 57

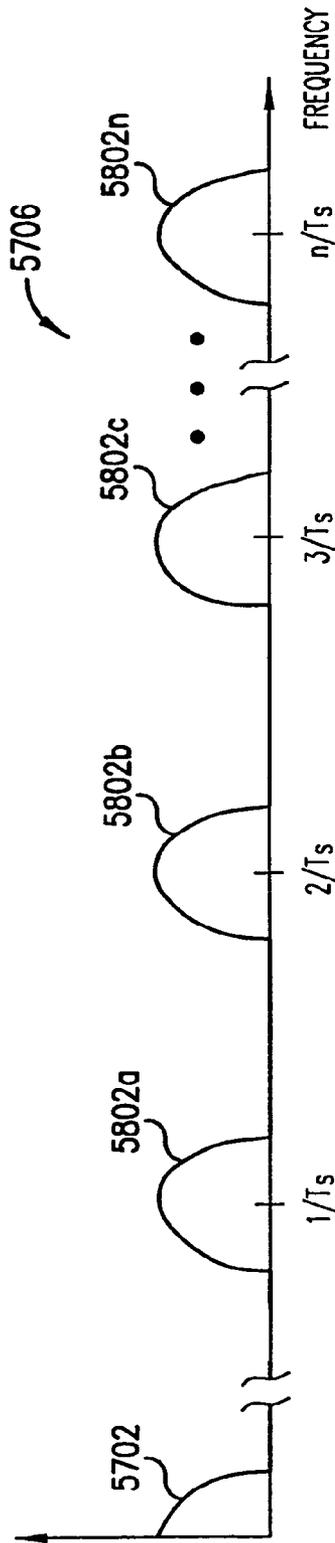


FIG. 58A

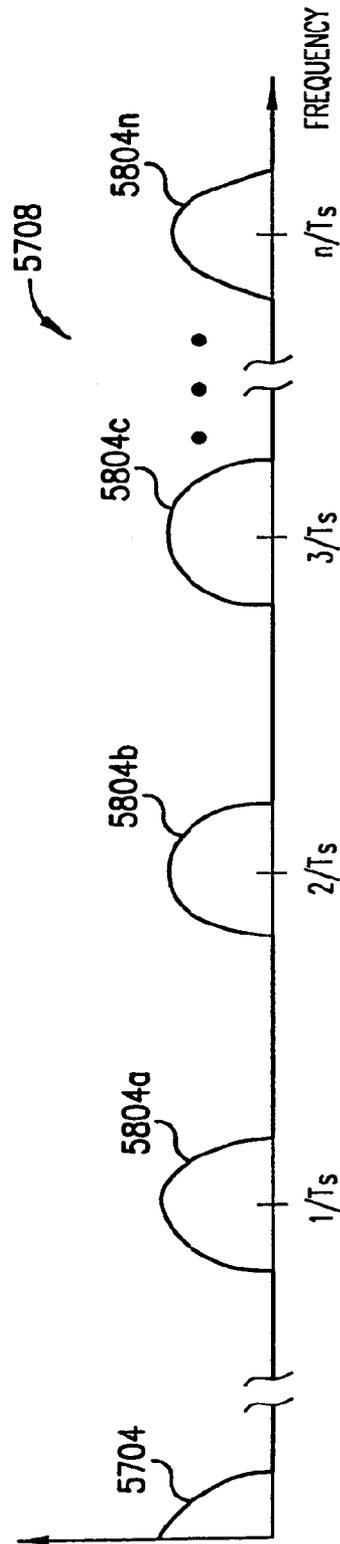


FIG. 58B

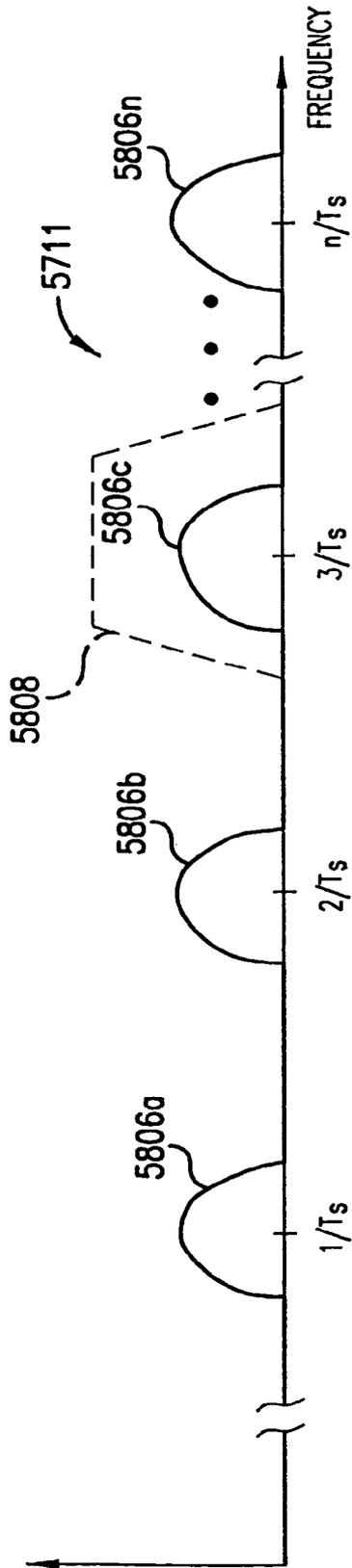


FIG. 58C

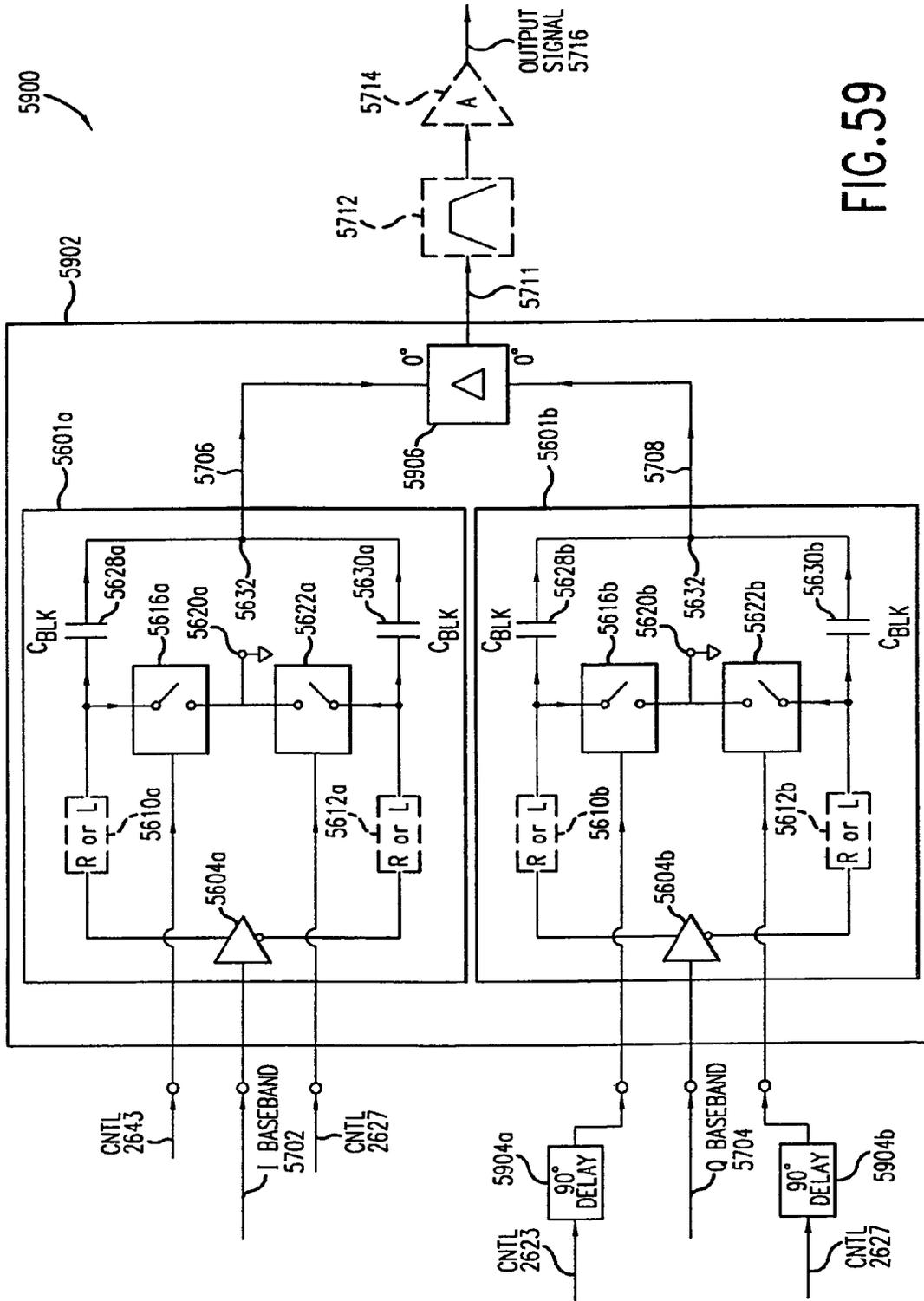


FIG. 59

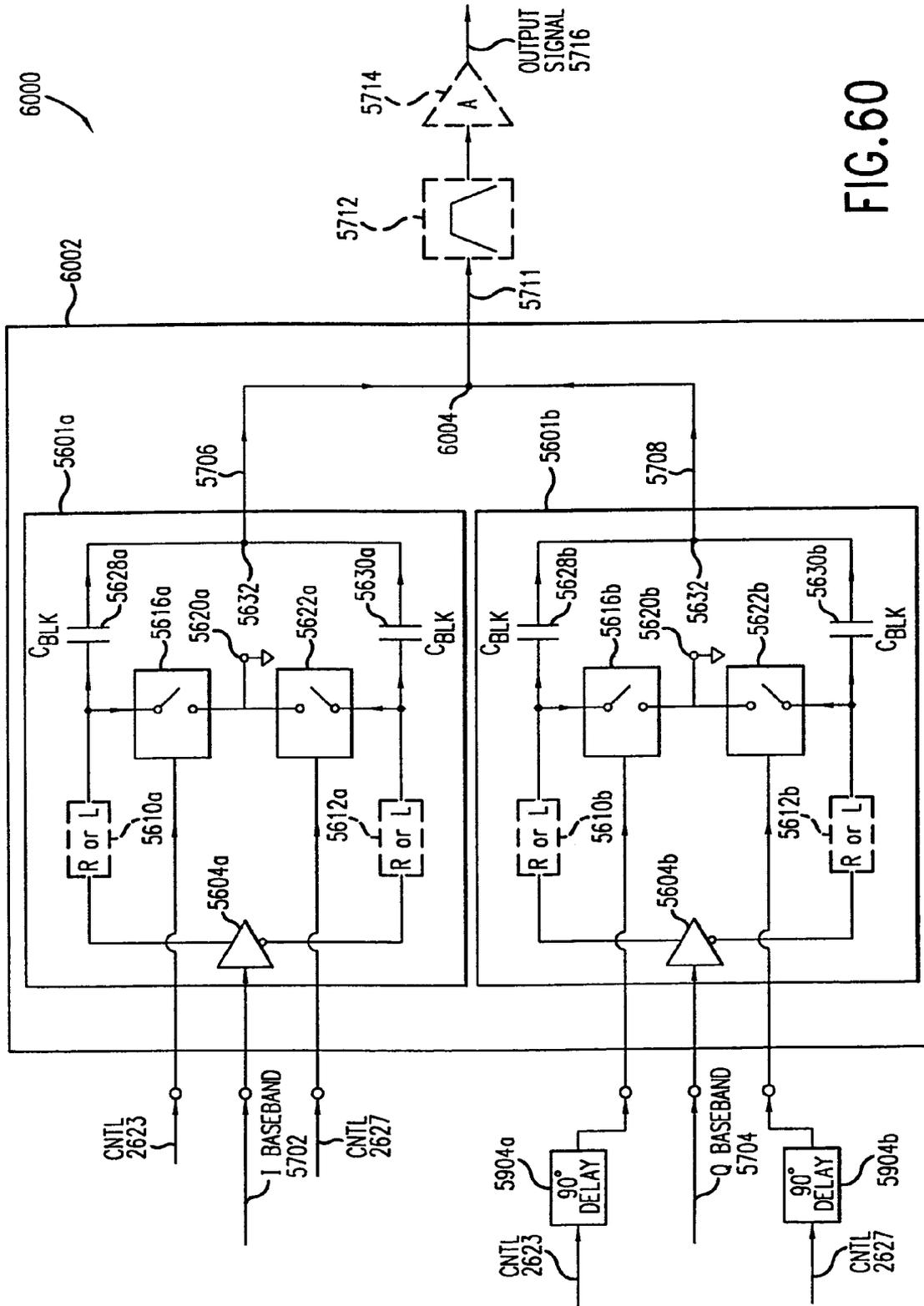


FIG. 60



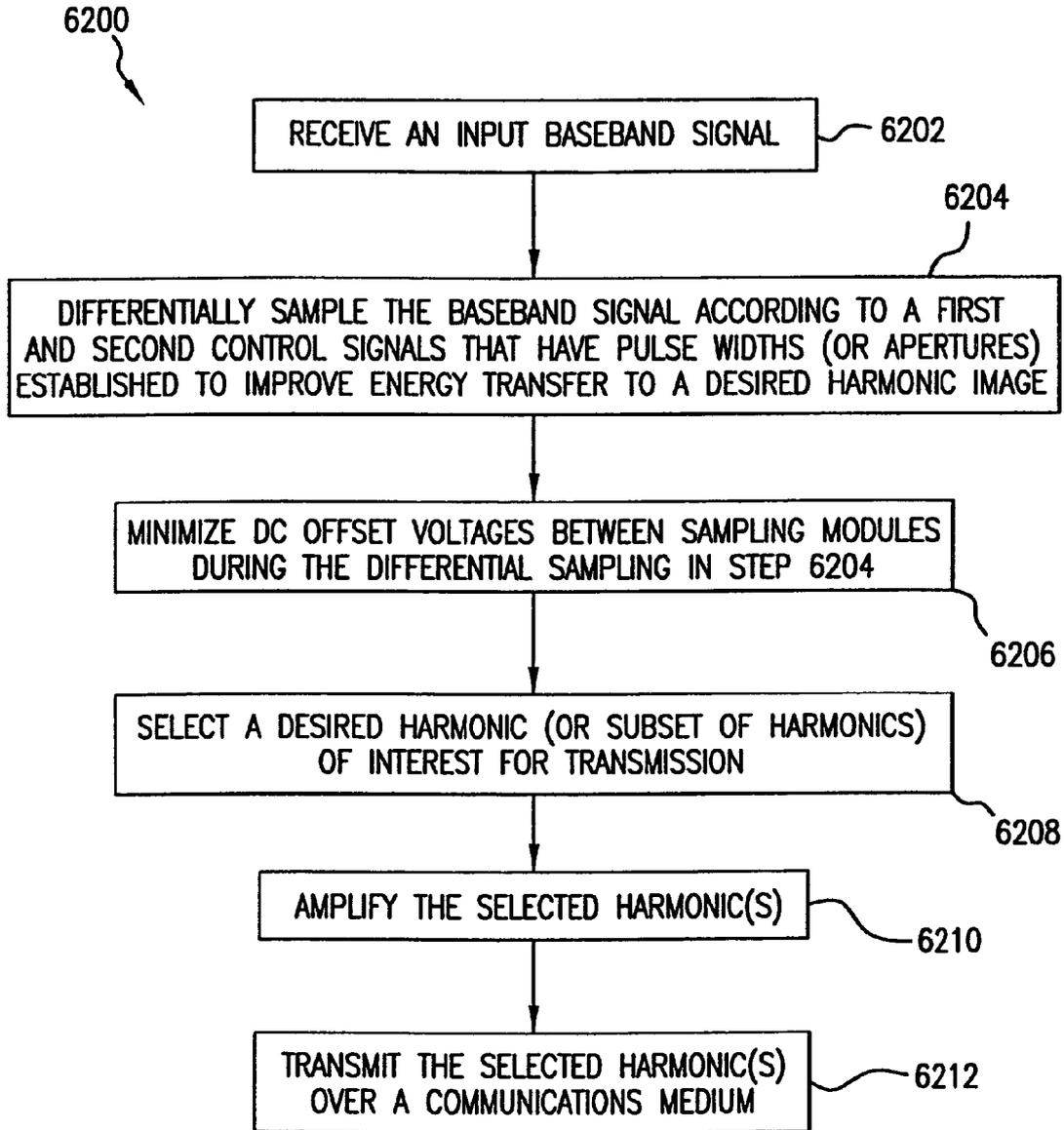


FIG.62

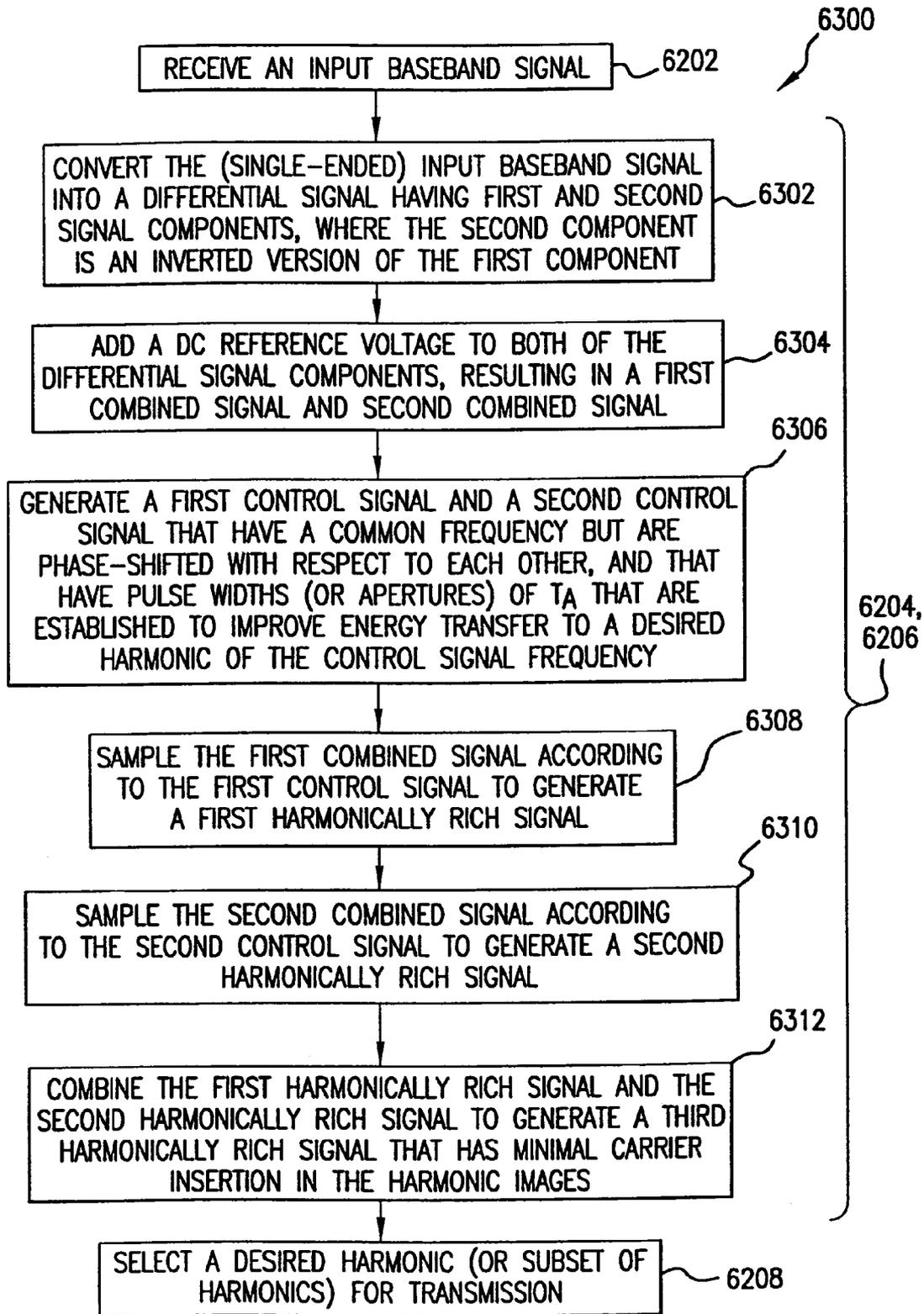


FIG.63

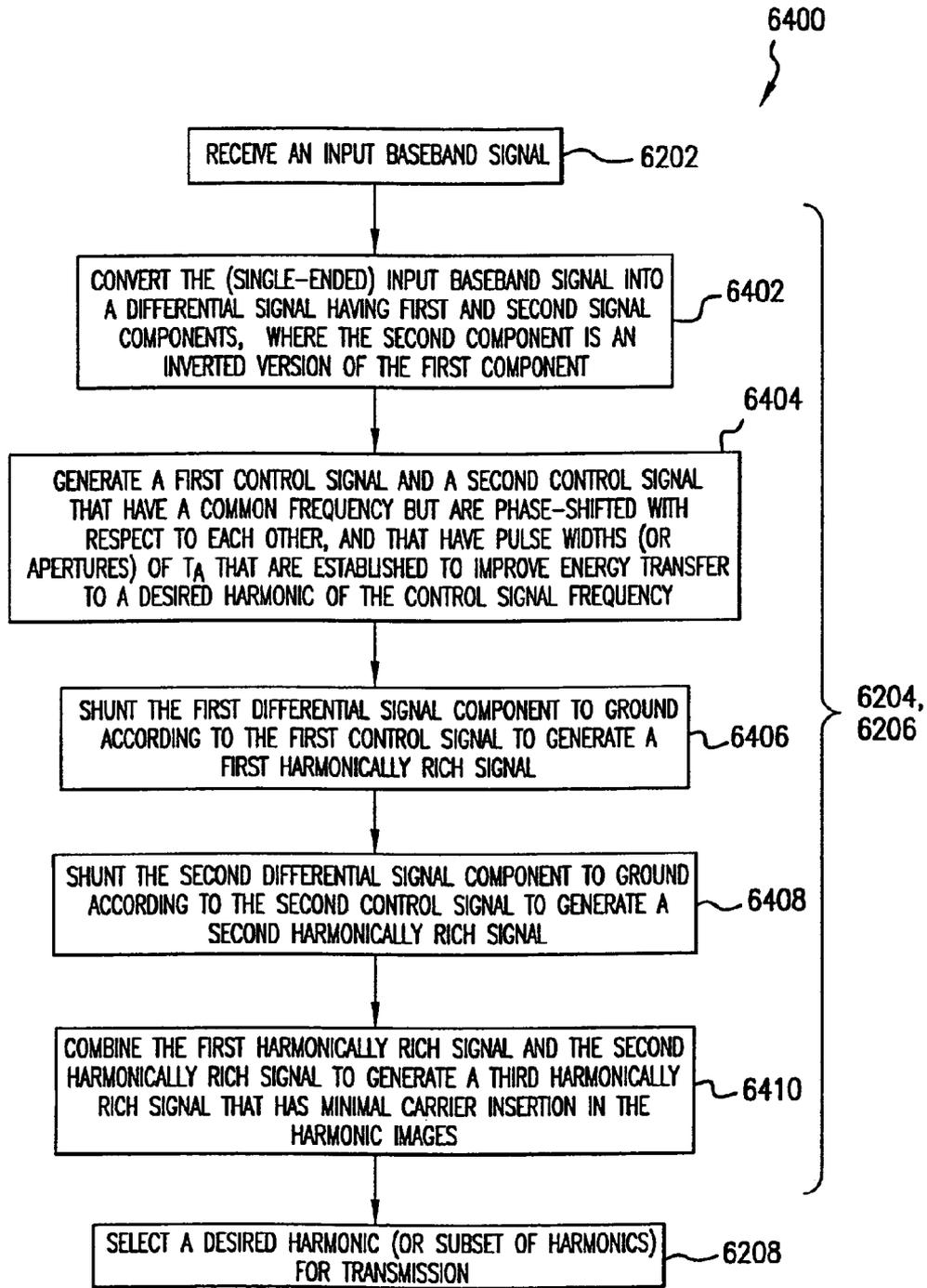


FIG.64

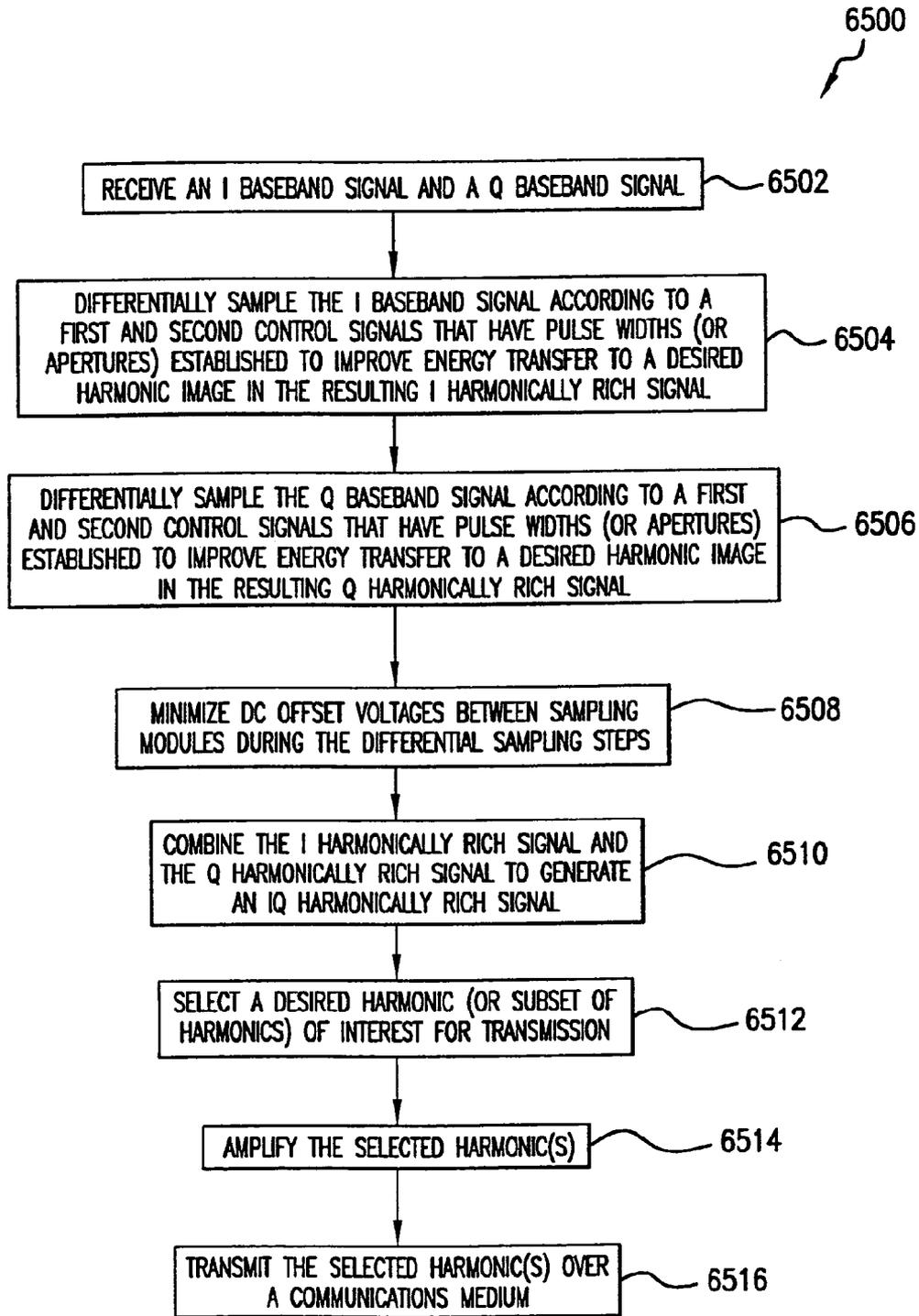


FIG.65

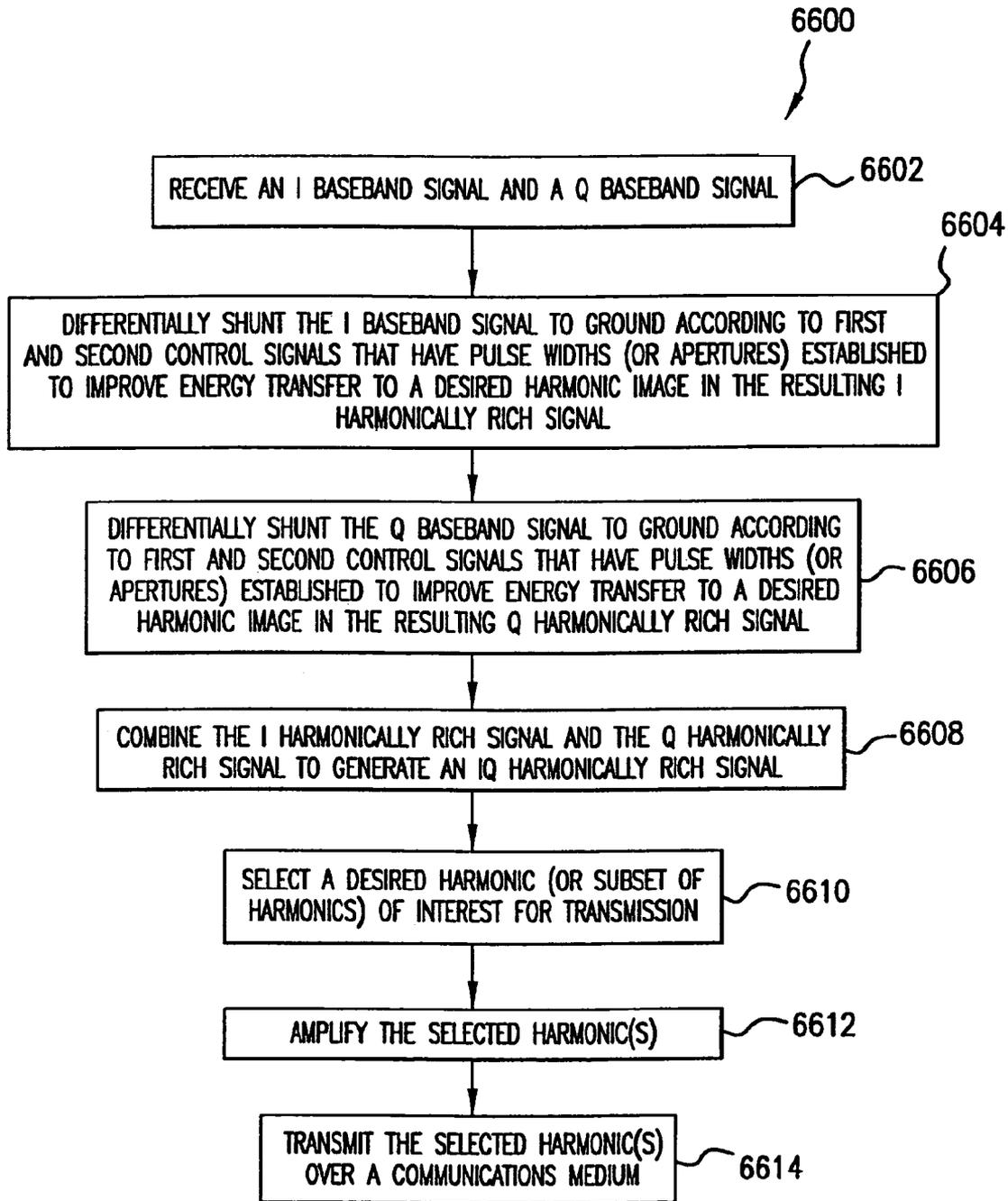


FIG.66

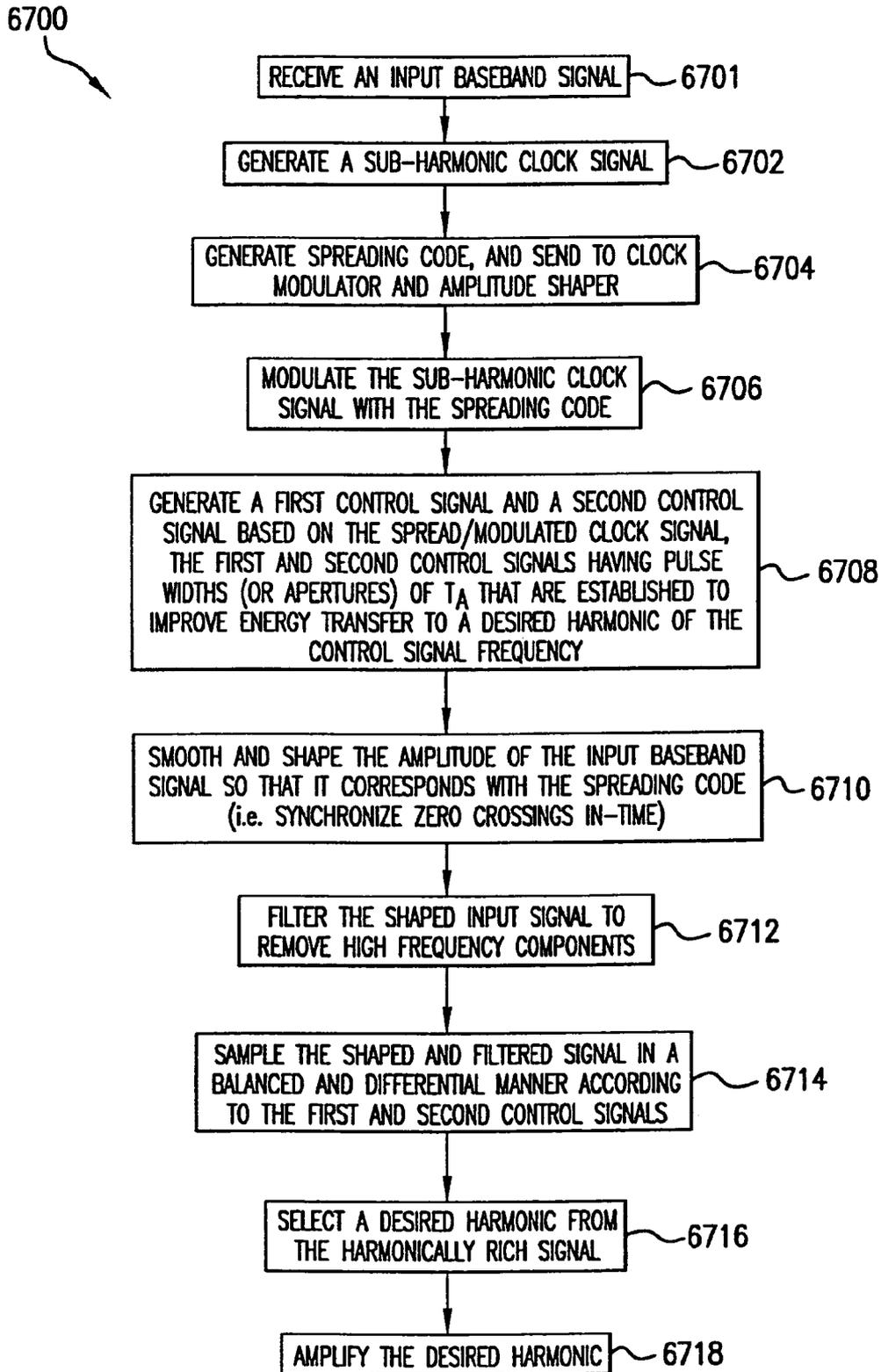


FIG.67

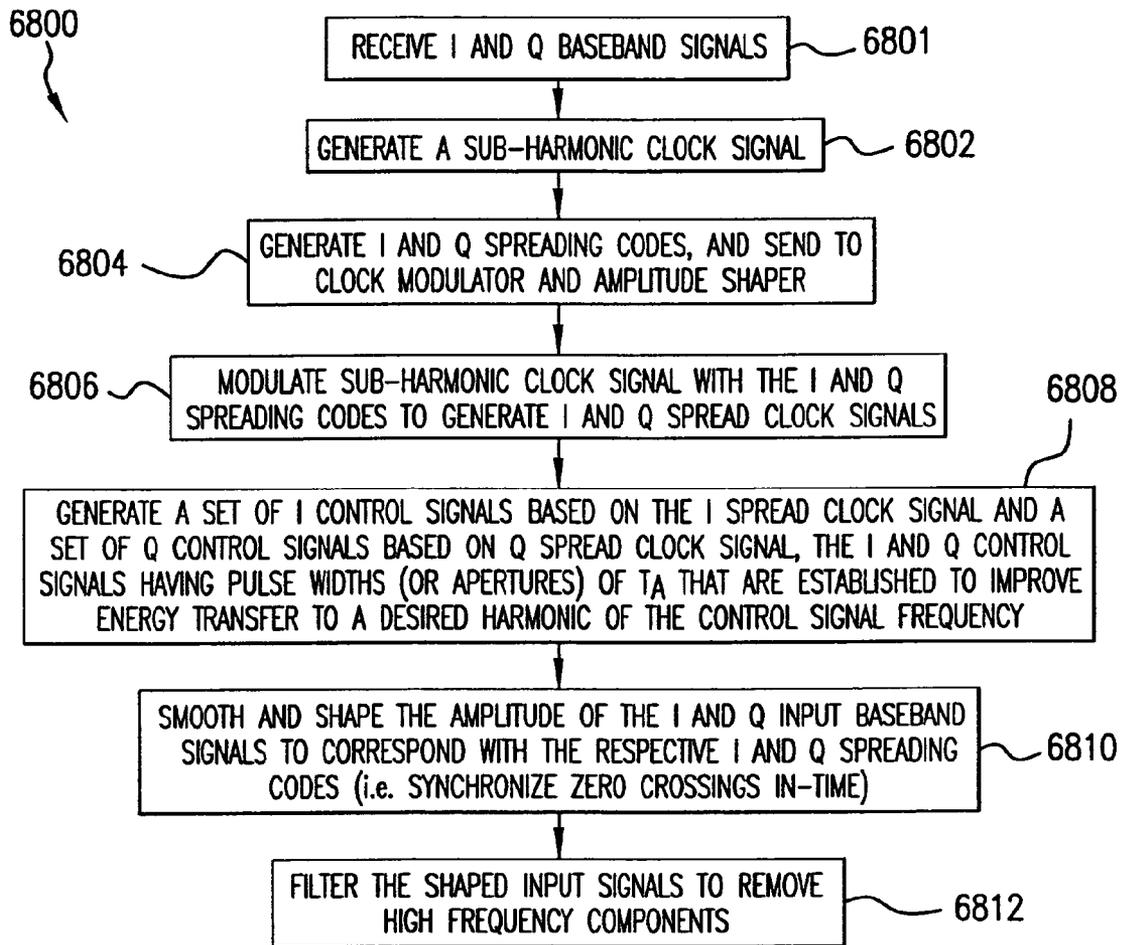


FIG.68A

6800  
(CONTINUED)

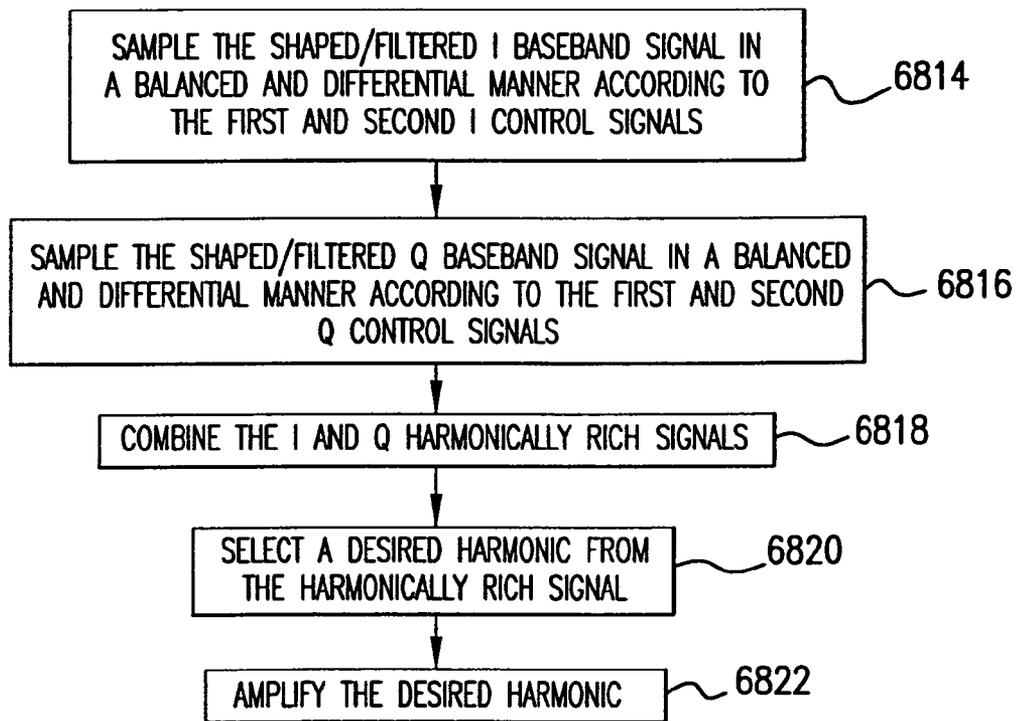


FIG. 68B

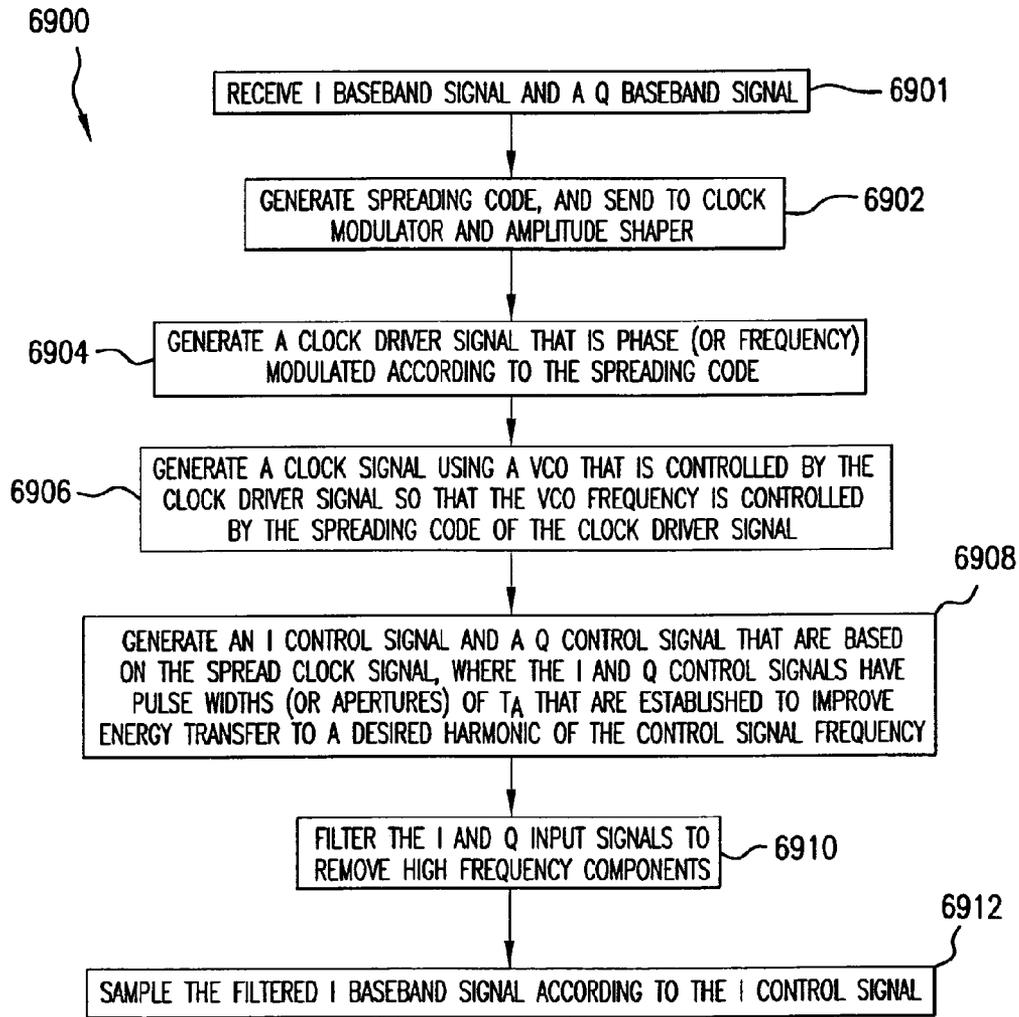


FIG. 69A

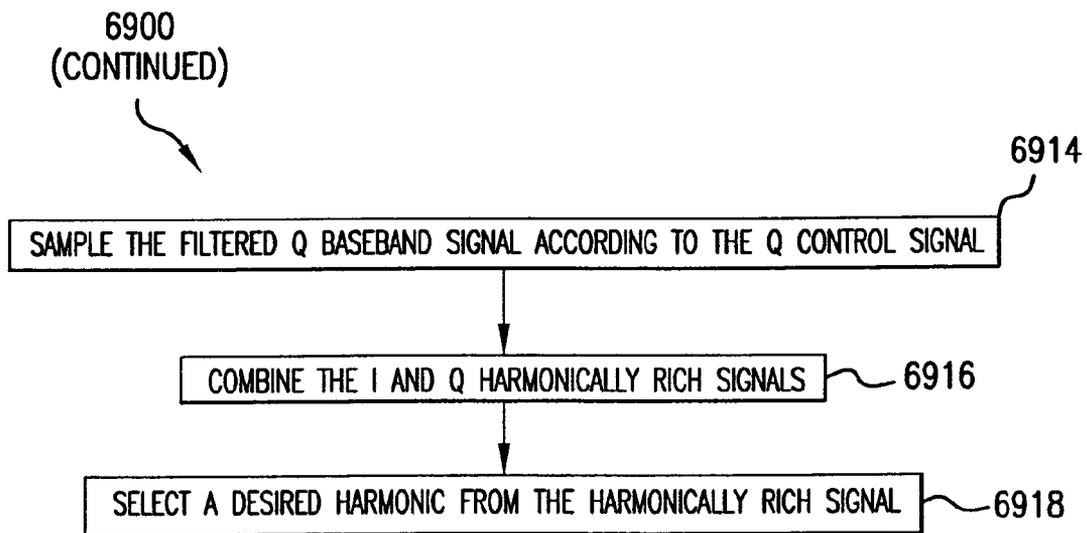


FIG.69B

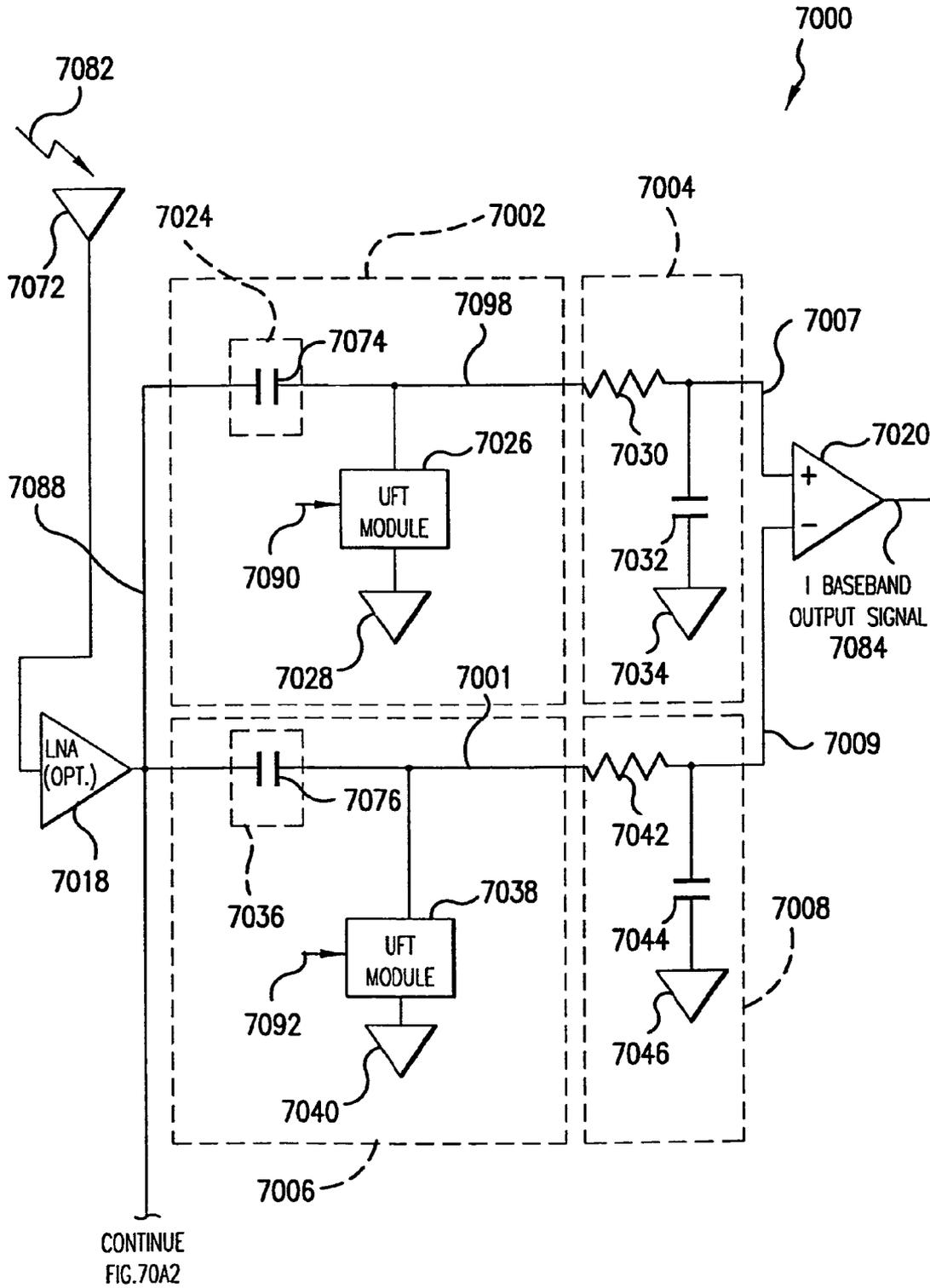


FIG. 70A1

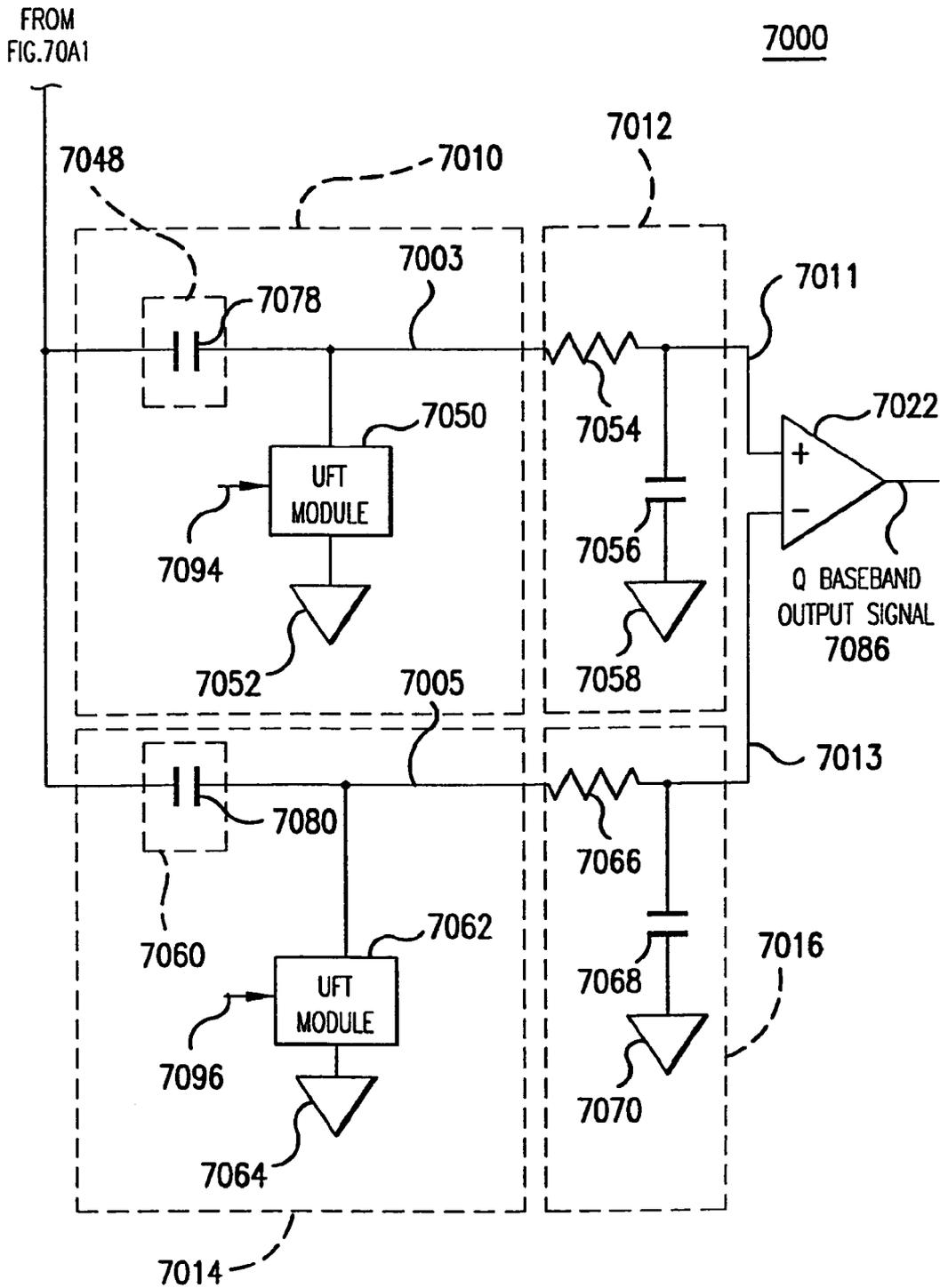


FIG. 70A2

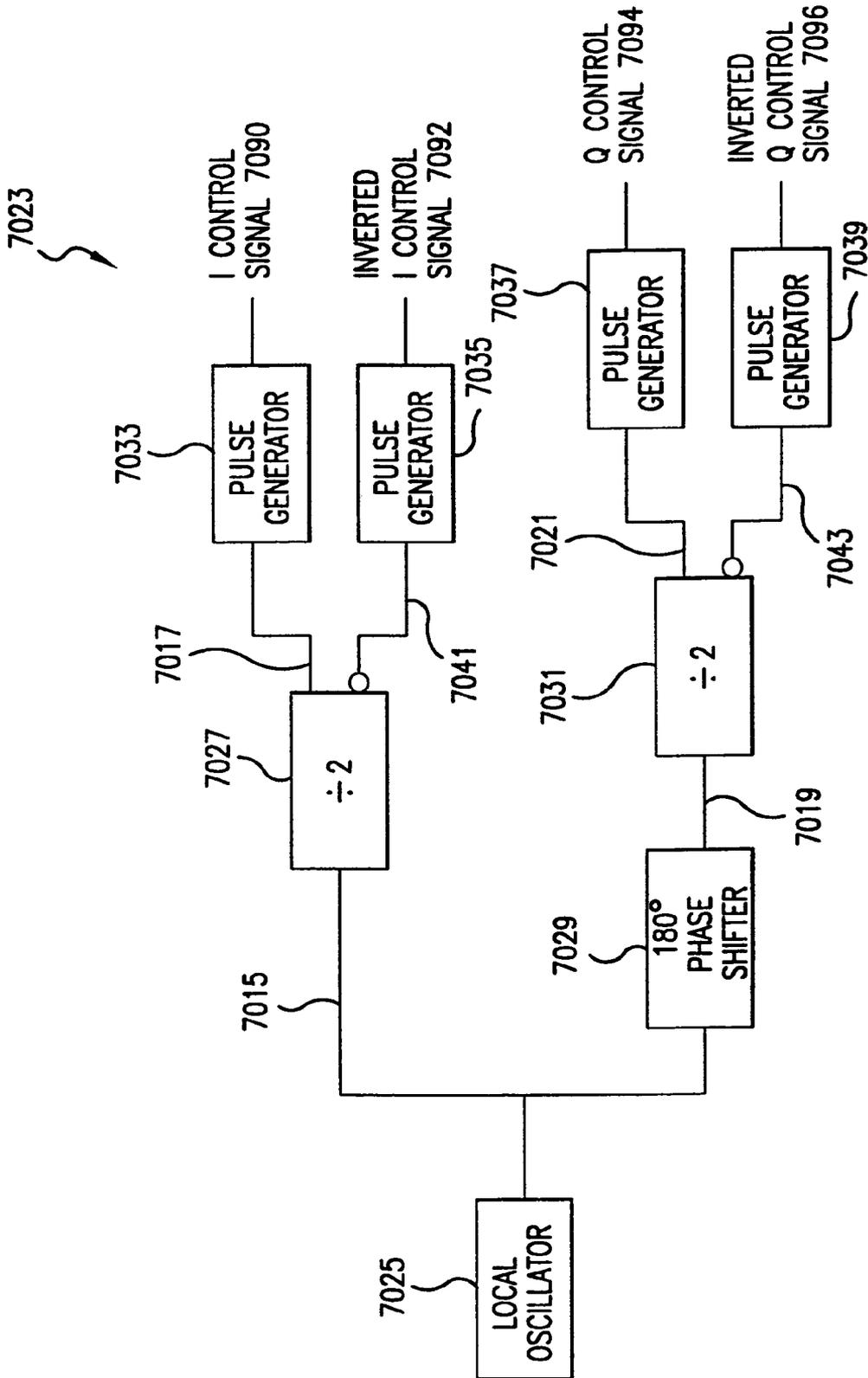


FIG. 700B

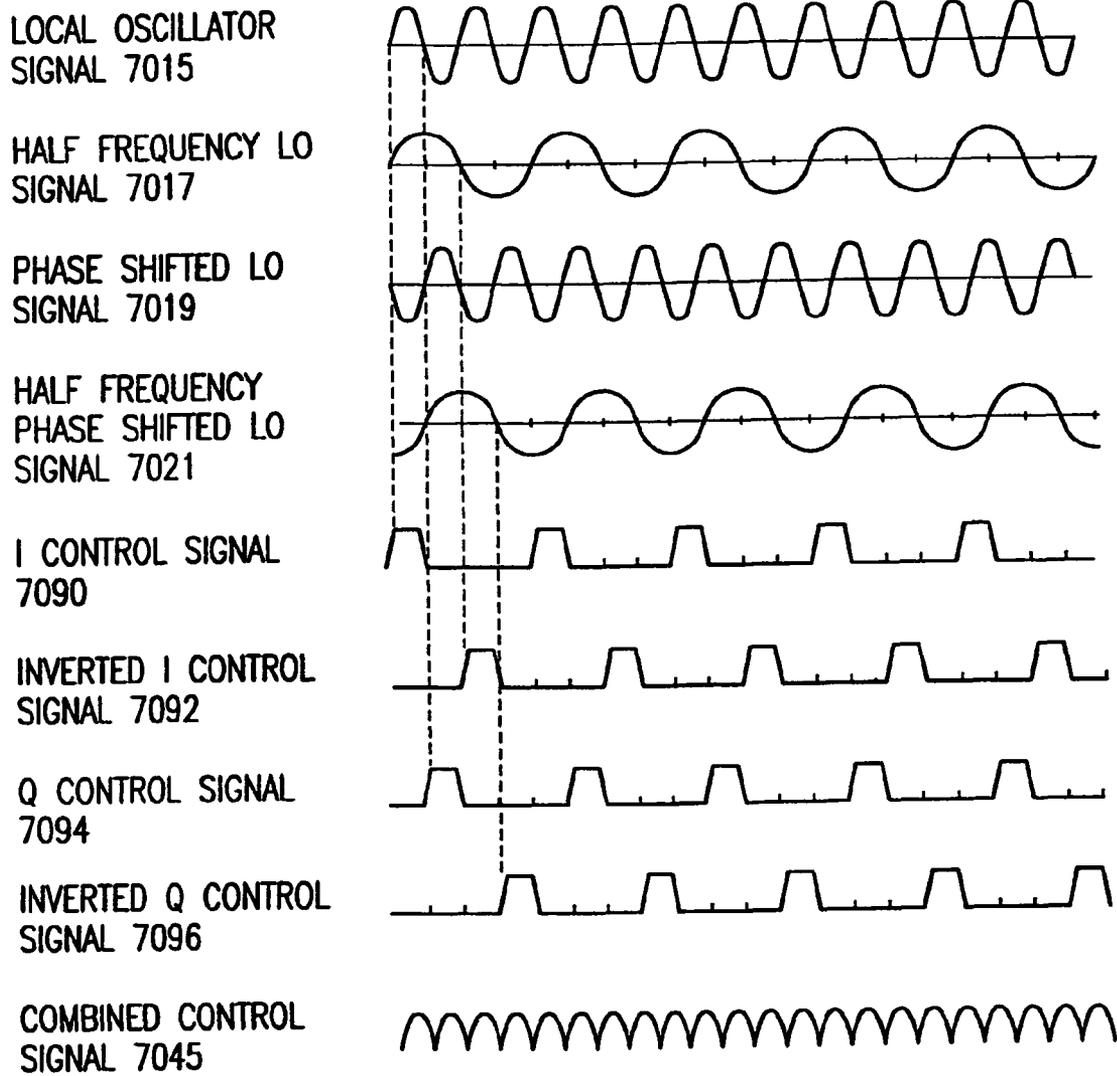
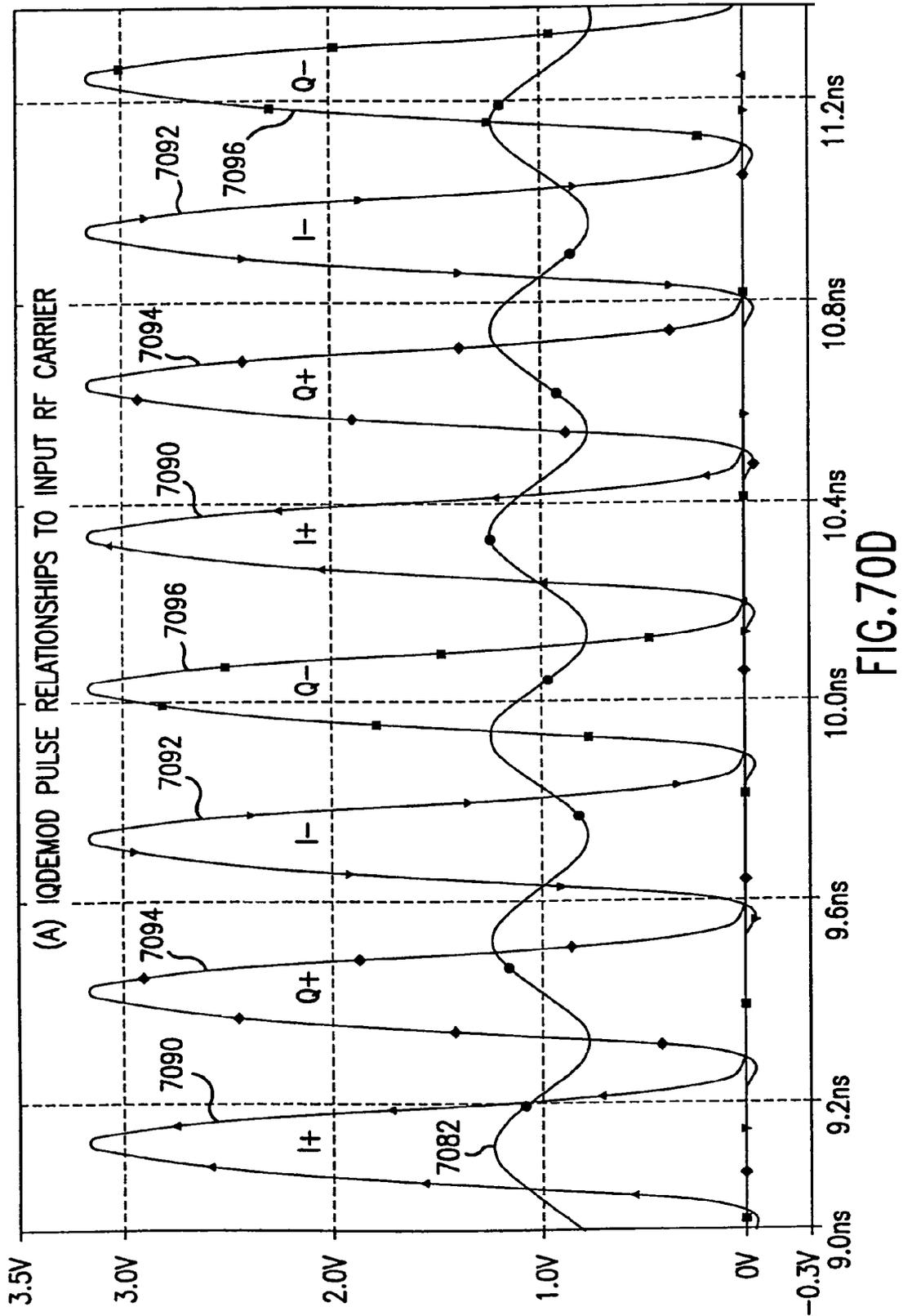


FIG.70C







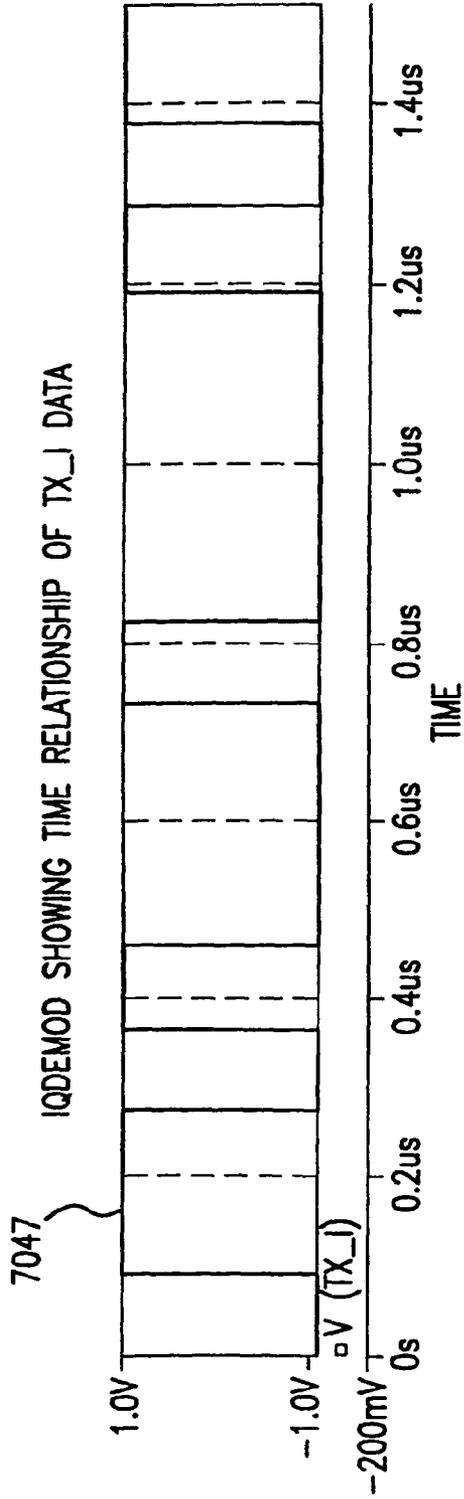


FIG.70F

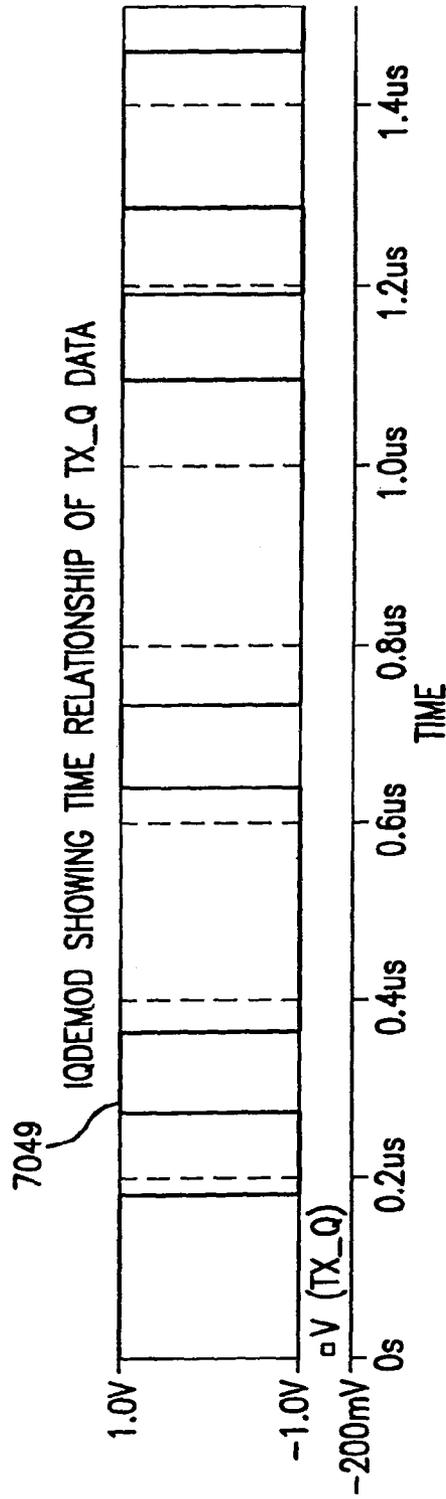


FIG.70G

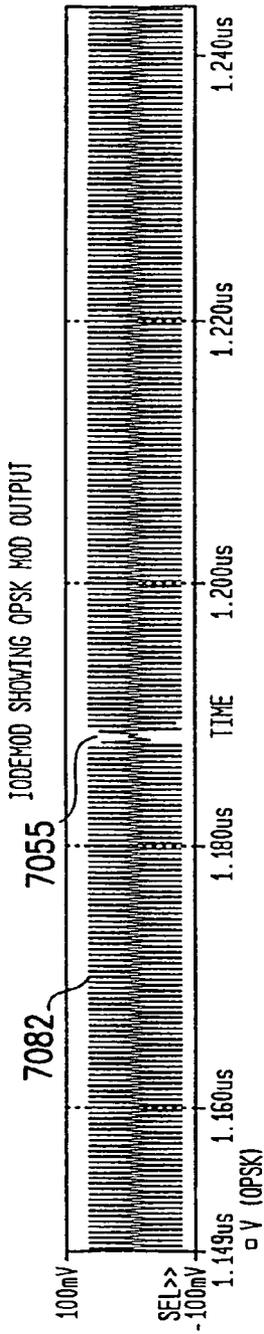


FIG. 70H

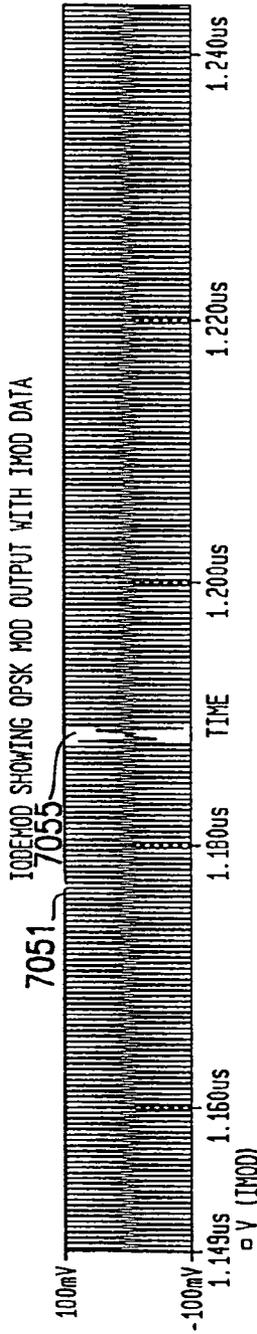


FIG. 70I

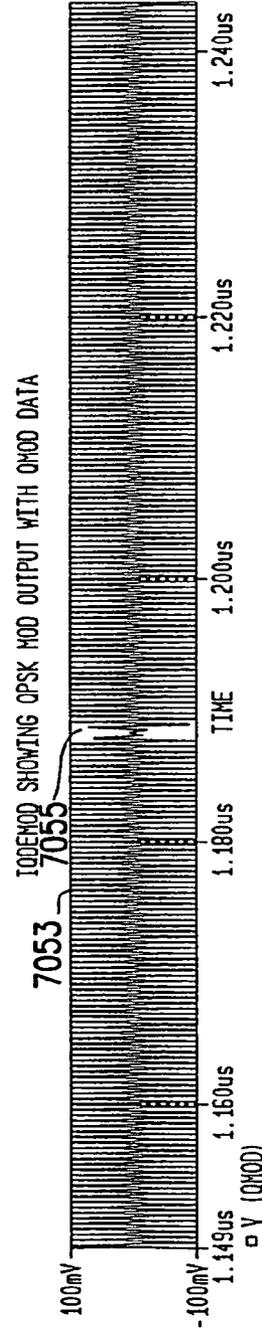


FIG. 70J

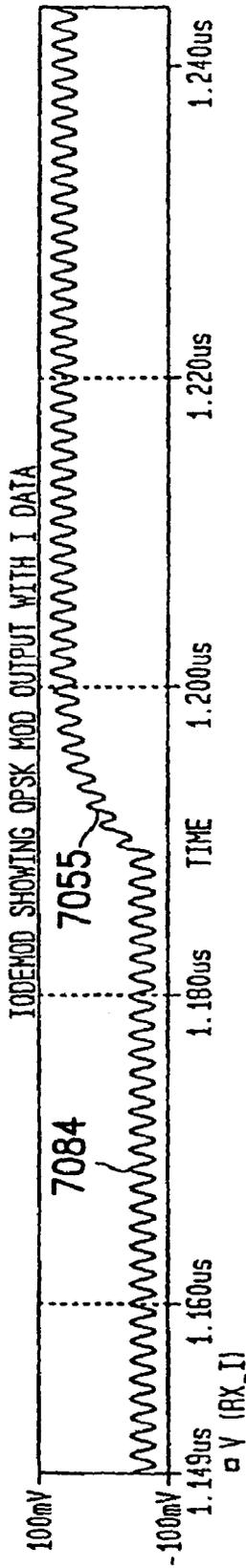


FIG. 70K

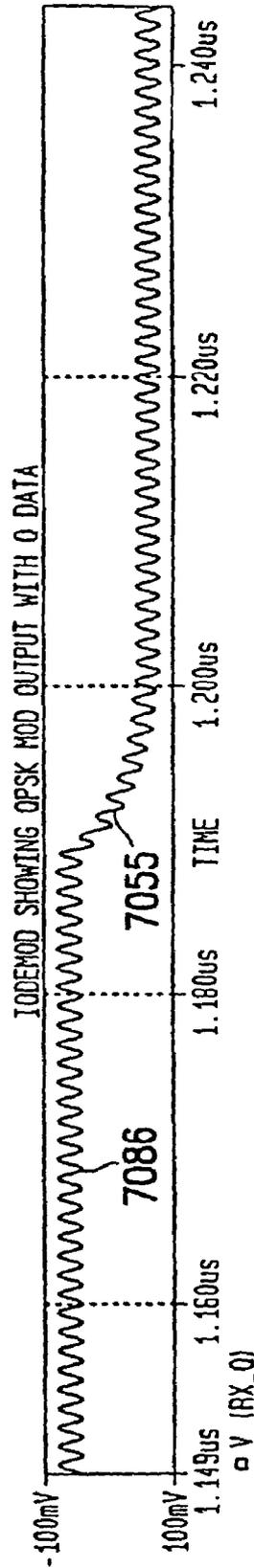


FIG. 70L

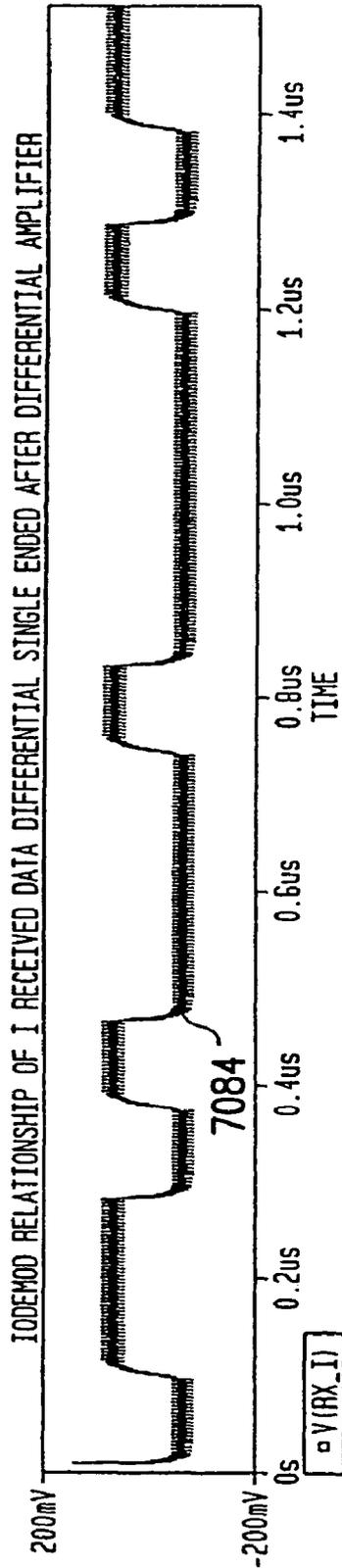


FIG. 70M

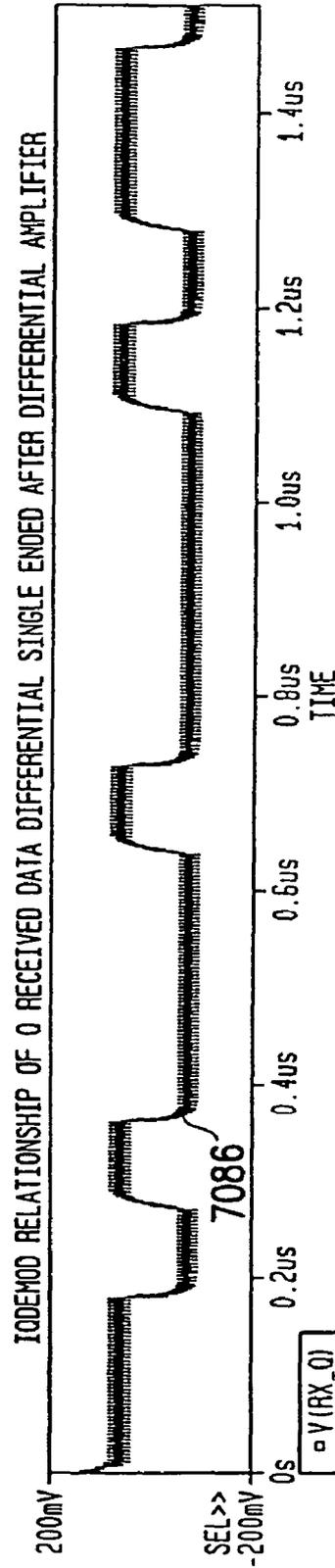


FIG. 70N

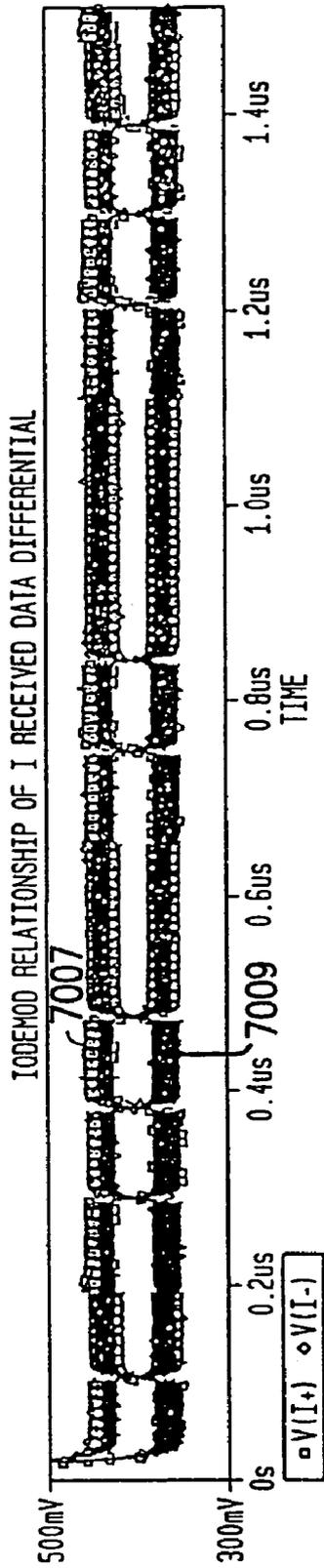


FIG. 700

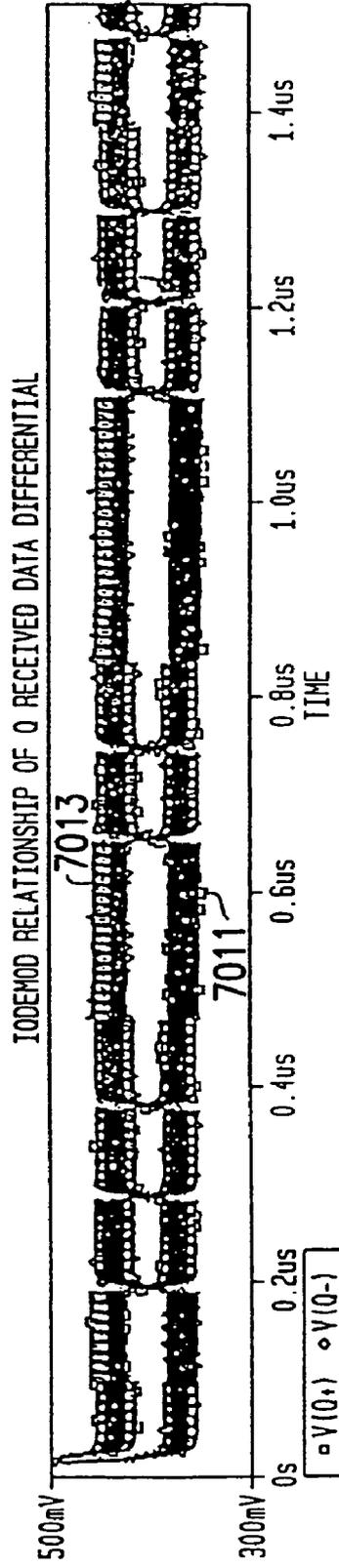


FIG. 70P

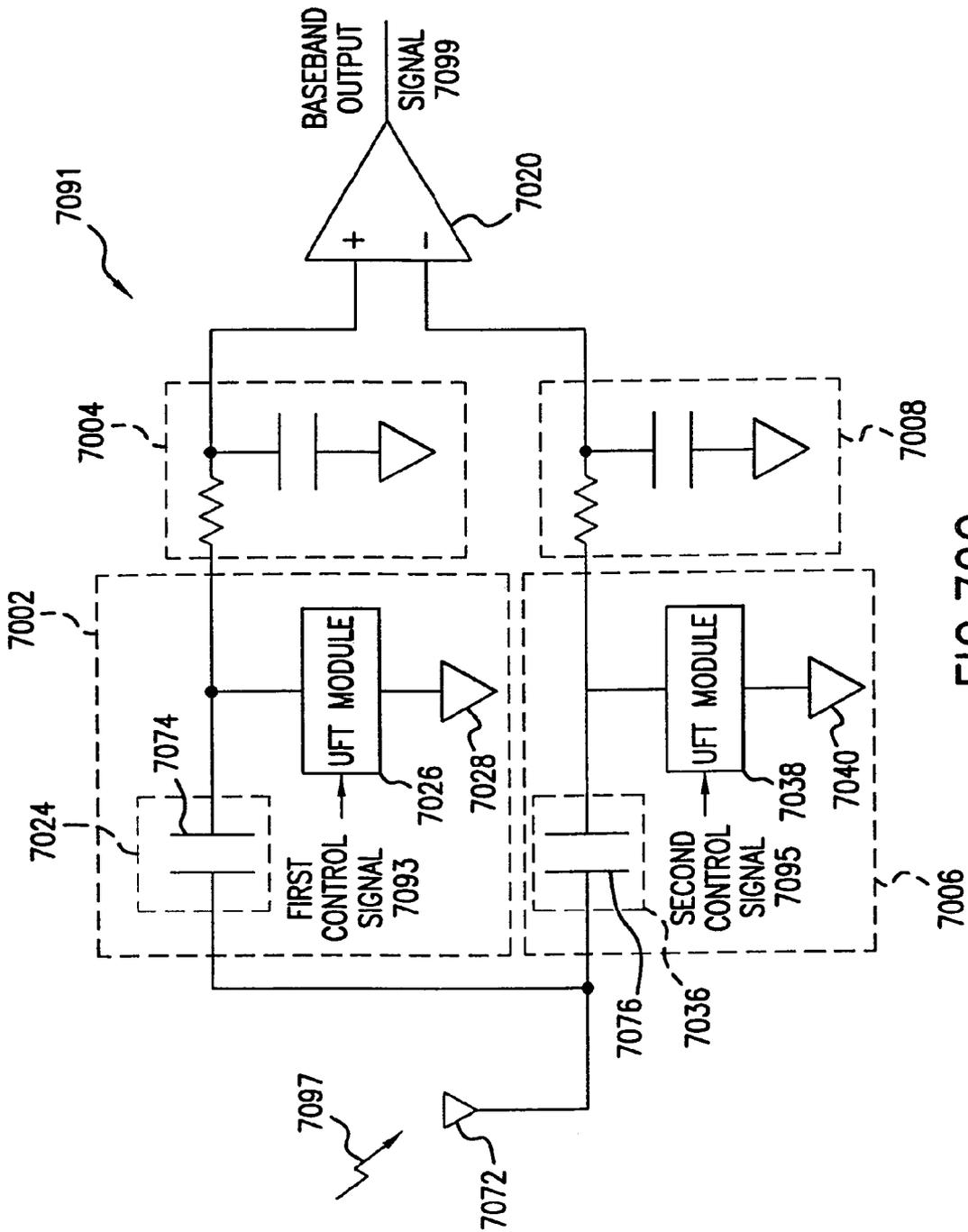


FIG. 700Q

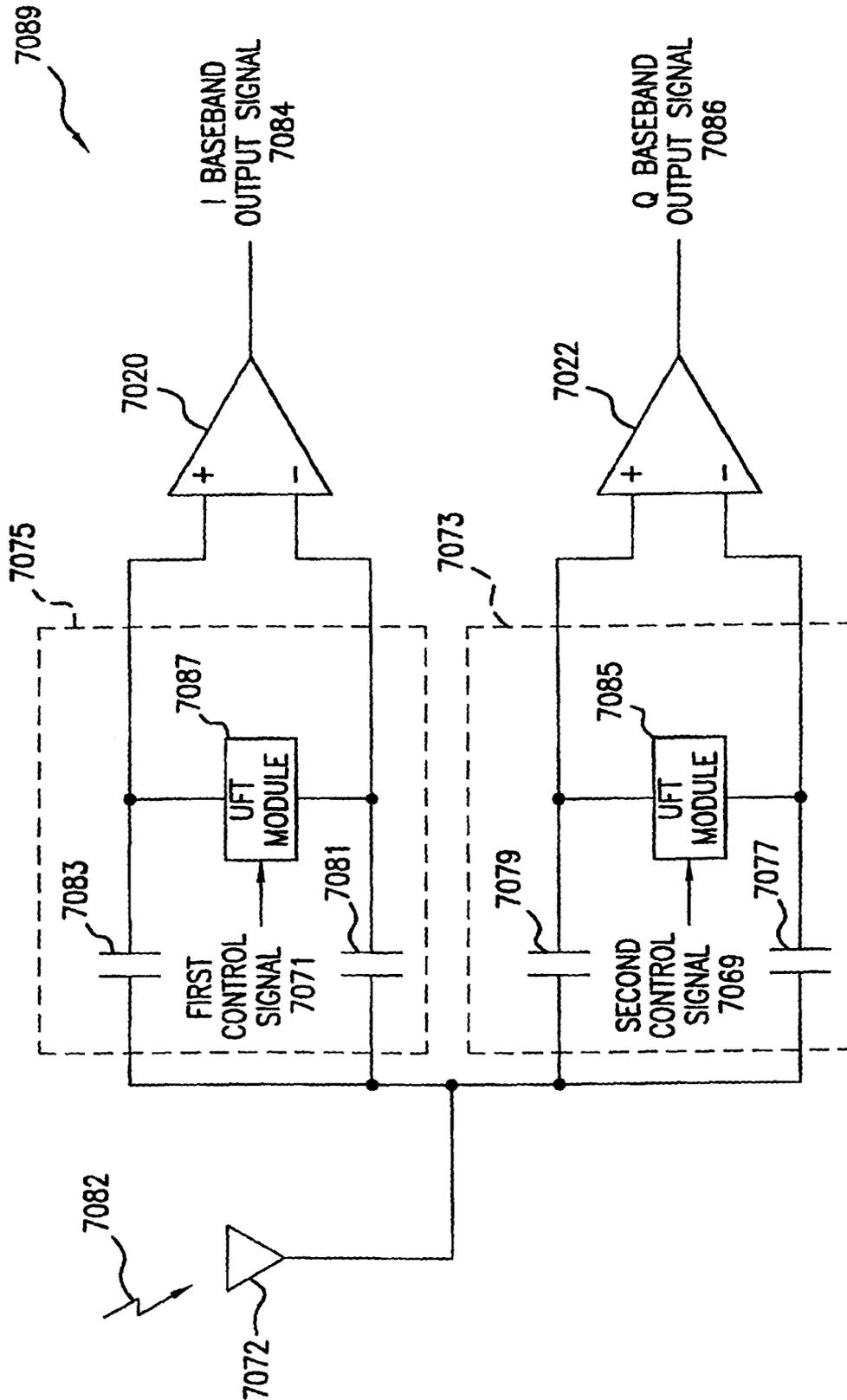


FIG. 70R

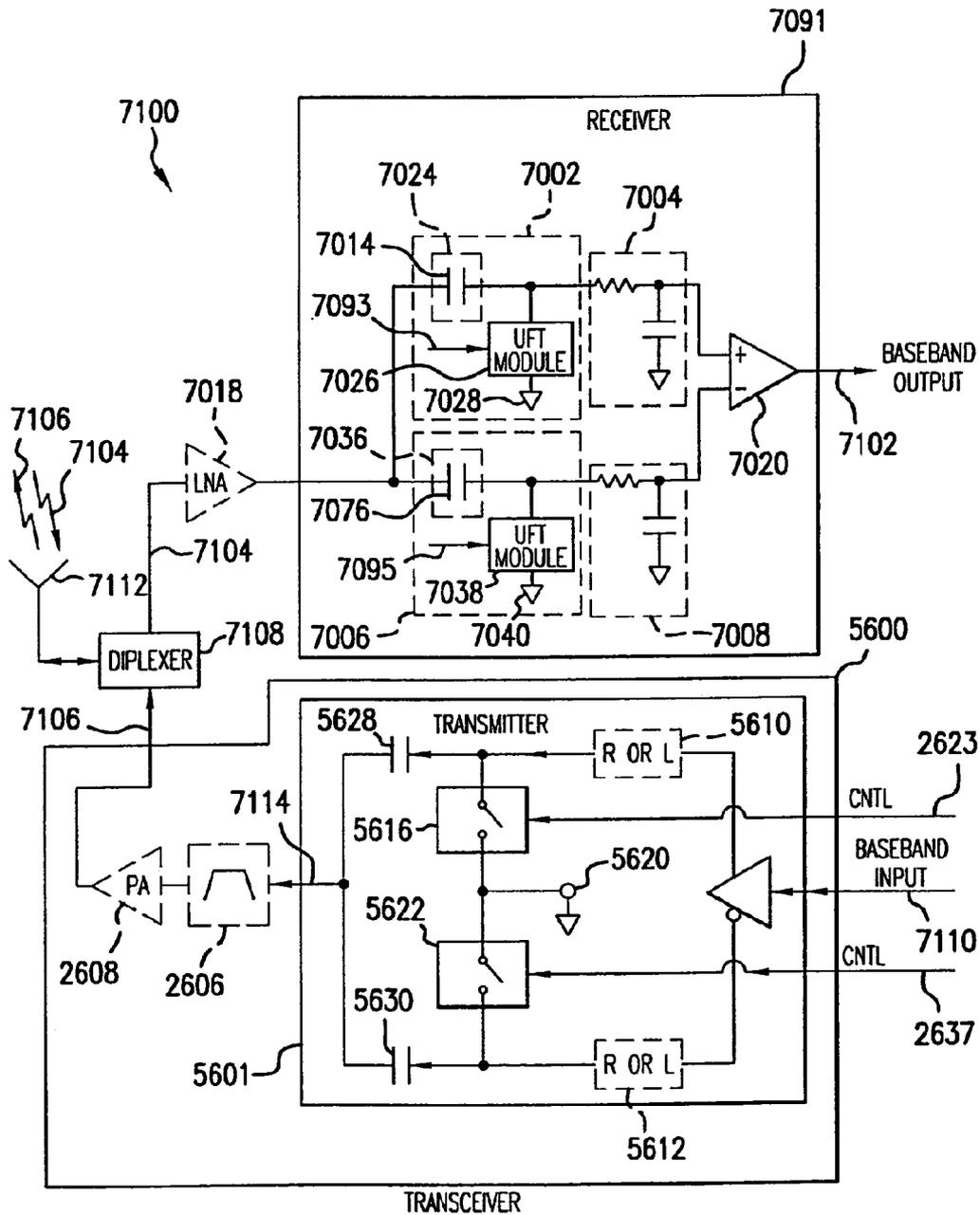


FIG.71

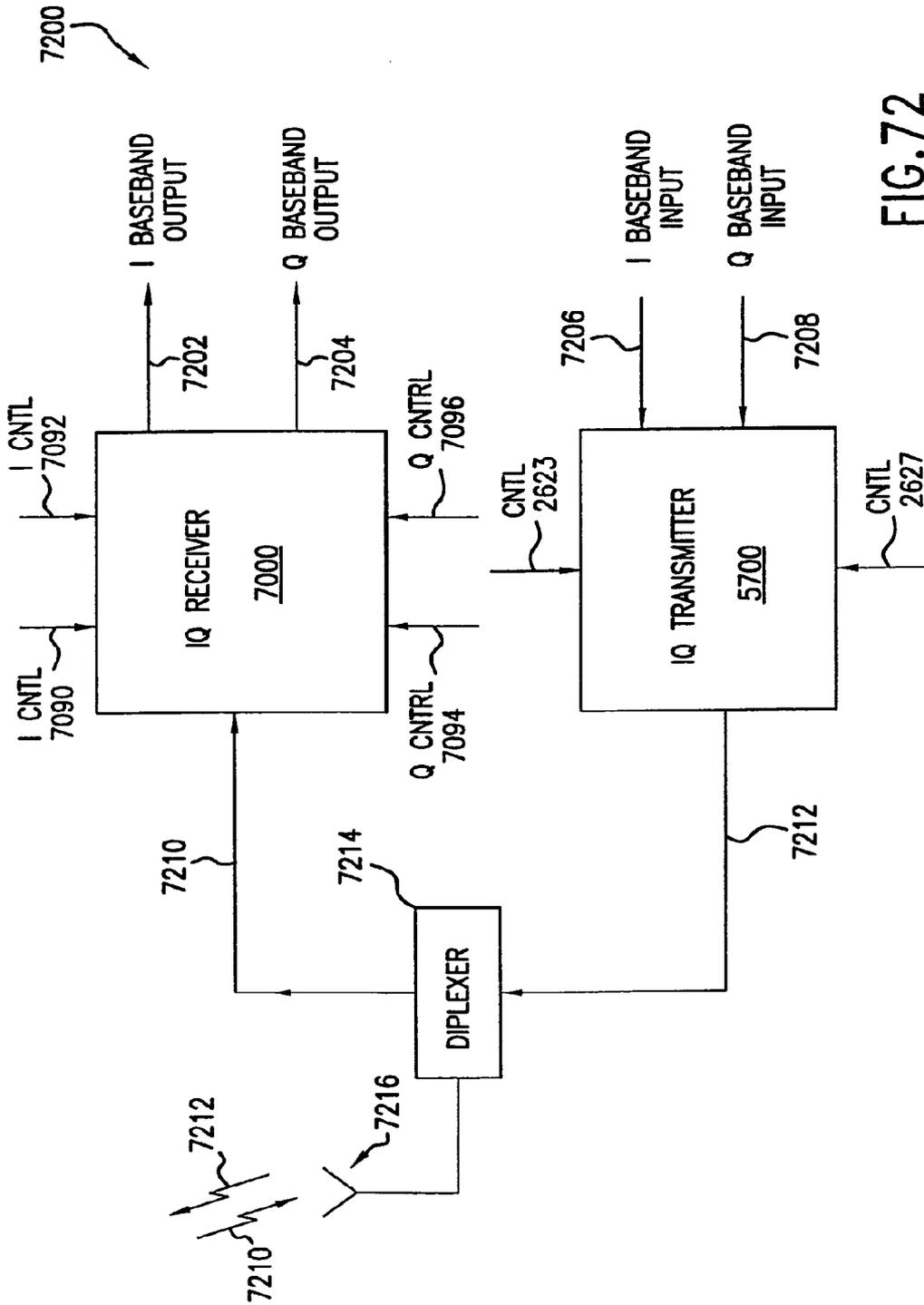


FIG. 72

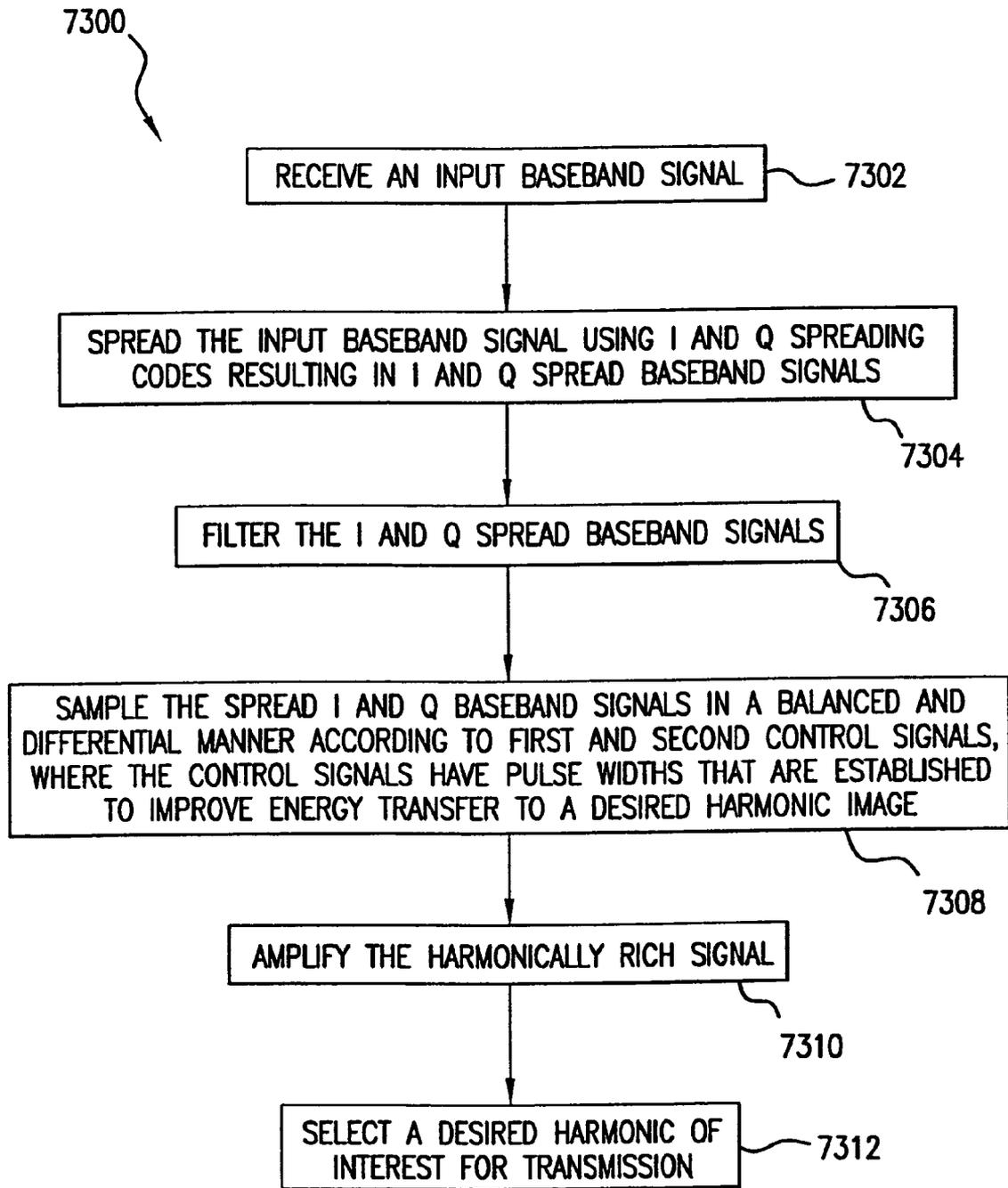


FIG.73

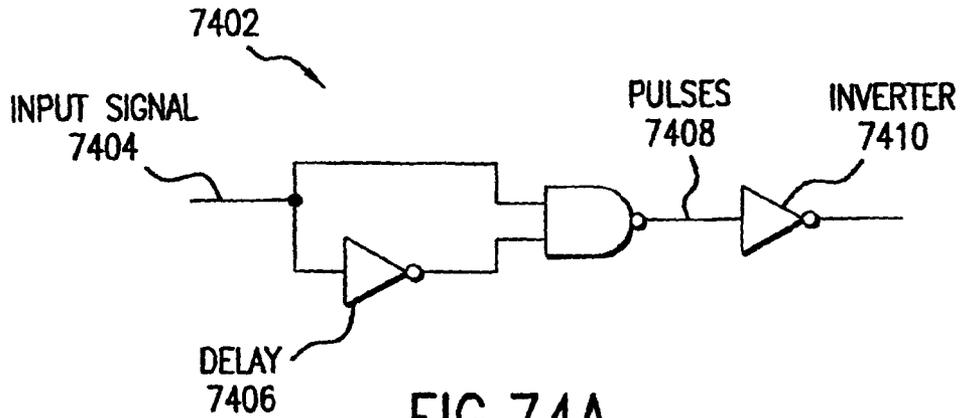


FIG. 74A

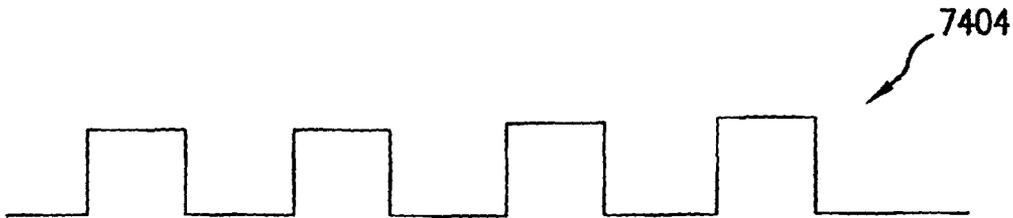


FIG. 74B

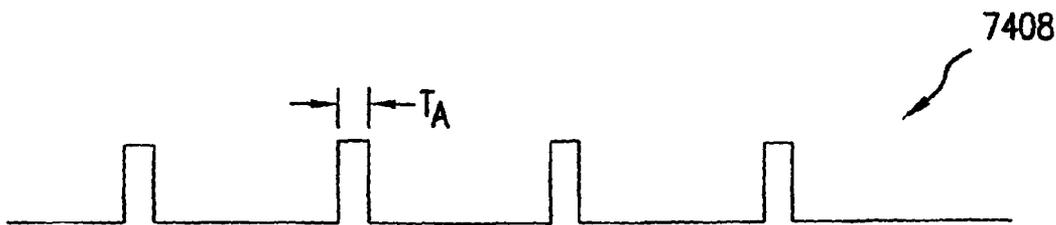


FIG. 74C

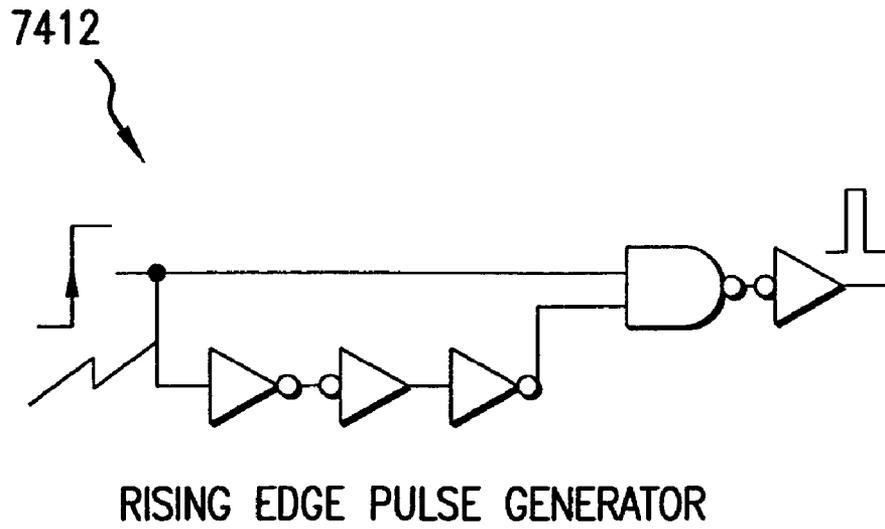


FIG.74D

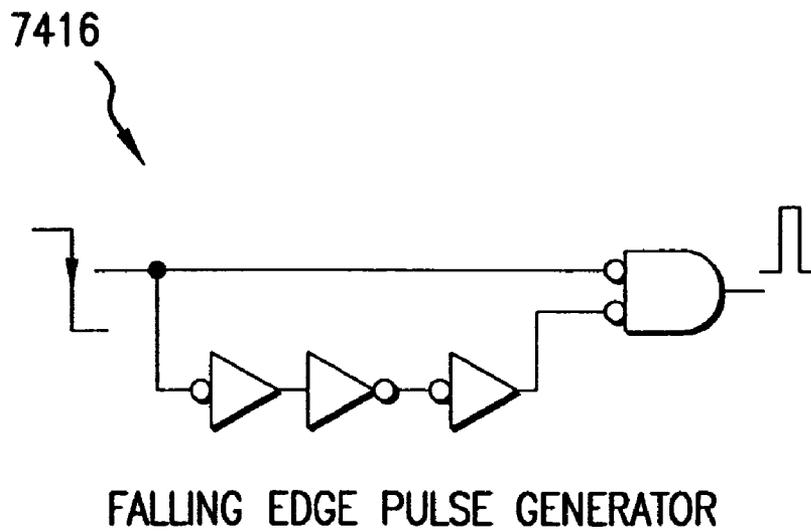


FIG.74E

**METHOD, SYSTEM AND APPARATUS FOR  
BALANCED FREQUENCY UP-CONVERSION  
OF A BASEBAND SIGNAL**

This application is a continuation to U.S. application Ser. No. 12/823,055, filed Jun. 24, 2010 now U.S. Pat. No. 8,077, 797, entitled "Method, System and Apparatus for Balanced Frequency Up-Conversion of a Baseband Signal" which is a continuation to U.S. patent application Ser. No. 11/015,653, filed Dec. 20, 2004 now U.S. Pat. No. 7,773,688, entitled "Method, System and Apparatus for Balanced Frequency Up-Conversion of a Baseband Signal." U.S. patent application Ser. No. 11/015,653 is a continuation of U.S. patent application Ser. No. 09/525,615 filed Mar. 14, 2000, which is now U.S. Pat. No. 6,853,690, which claims benefit of the following: U.S. Provisional Application No. 60/177,381, filed on Jan. 24, 2000; U.S. Provisional Application No. 60/171,502, filed Dec. 22, 1999; U.S. Provisional Application No. 60/177,705, filed on Jan. 24, 2000; U.S. Provisional Application No. 60/129,839, filed on Apr. 16, 1999; U.S. Provisional Application No. 60/158,047, filed on Oct. 7, 1999; U.S. Provisional Application No. 60/171,349, filed on Dec. 21, 1999; U.S. Provisional Application No. 60/177,702, filed on Jan. 24, 2000; U.S. Provisional Application No. 60/180,667, filed on Feb. 7, 2000; and U.S. Provisional Application No. 60/171,496, filed on Dec. 22, 1999. The subject matter of all of the above-referenced applications is incorporated herein by reference as if fully set forth herein.

**CROSS-REFERENCE TO OTHER  
APPLICATIONS**

The following applications of common assignee are related to the present application, and are herein incorporated by reference in their entireties:

"Method and System for Down-Converting Electromagnetic Signals," Ser. No. 09/176,022, filed Oct. 21, 1998;

"Method and System for Frequency Up-Conversion," Ser. No. 09/176,154, filed Oct. 21, 1998;

"Method and System for Ensuring Reception of a Communications Signal," Ser. No. 09/176,415, filed Oct. 21, 1998;

"Integrated Frequency Translation And Selectivity," Ser. No. 09/175,966, filed Oct. 21, 1998;

"Universal Frequency Translation, and Applications of Same," Ser. No. 09/176,027, filed Oct. 21, 1998;

"Applications of Universal Frequency Translation," filed Mar. 3, 1999, Ser. No. 09/261,129, filed Mar. 3, 1999;

"Matched Filter Characterization and Implementation of Universal Frequency Translation Method and Apparatus," Ser. No. 09/521,878, filed Mar. 9, 2000;

"Spread Spectrum Applications of Universal Frequency Translation," Ser. No. 09/525,185; and

"DC Offset, Re-radiation, and I/Q Solutions Using Universal Frequency Translation Technology," Ser. No. 11/059,536.

**BACKGROUND OF THE INVENTION**

**1. Field of the Invention**

The present invention is generally related to frequency up-conversion of a baseband signal, and applications of same. The invention is also directed to embodiments for frequency down-conversion, and to transceivers.

**2. Related Art**

Various communication components and systems exist for performing frequency up-conversion and down-conversion of electromagnetic signals.

**SUMMARY OF THE INVENTION**

The present invention is related to up-converting a baseband signal, and applications of same. Such applications include, but are not limited to, up-converting a spread spectrum signal directly from baseband to radio frequency (RF) without utilizing any intermediate frequency (IF) processing. The invention is also related to frequency down-conversion.

In embodiments, the invention differentially samples a baseband signal according to first and second control signals, resulting in a harmonically rich signal. The harmonically rich signal contains multiple harmonic images that each contain the necessary amplitude, frequency, and/or phase information to reconstruct the baseband signal. The harmonic images in the harmonically rich signal repeat at the harmonics of the sampling frequency ( $1/T_s$ ) that are associated with the first and second control signals. In other words, the sampling is performed sub-harmonically according to the control signals. Additionally, the control signals include pulses that have an associated pulse width  $T_A$  that is established to improve energy transfer to a desired harmonic image in the harmonically rich signal. The desired harmonic image can optionally be selected using a bandpass filter for transmission over a communications medium.

In operation, the invention converts the input baseband signal from a (single-ended) input into a differential baseband signal having first and second components. The first differential component is substantially similar to the input baseband signal, and the second differential component is an inverted version of the input baseband signal. The first differential component is sampled according to the first control signal, resulting in a first harmonically rich signal. Likewise, the second differential component is sampled according to the second control signal, resulting in a second harmonically rich signal. The first and second harmonically rich signals are combined to generate the output harmonically rich signal.

The sampling modules that perform the differentially sampling can be configured in a series or shunt configuration. In the series configuration, the baseband input is received at one port of the sampling module, and is gated to a second port of the sampling module, to generate the harmonically rich signal at the second port of the sampling module. In the shunt configuration, the baseband input is received at one port of the sampling module and is periodically shunted to ground at the second port of the sampling module, according to the control signal. Therefore, in the shunt configuration, the harmonically rich signal is generated at the first port of the sampling module and coexists with the baseband input signal at the first port.

The first control signal and second control signals that control the sampling process are phase shifted relative to one another. In embodiments of the invention, the phase-shift is 180 degree in reference to a master clock signal, although the invention includes other phase shift values. Therefore, the sampling modules alternately sample the differential components of the baseband signal. Additionally as mentioned above, the first and second control signals include pulses having a pulse width  $T_A$  that is established to improve energy transfer to a desired harmonic in the harmonically rich signal during the sampling process. More specifically, the pulse width  $T_A$  is a non-negligible fraction of a period associated with a desired harmonic of interest. In an embodiment, the pulse width  $T_A$  is one-half of a period of the harmonic of interest. Additionally, in an embodiment, the frequency of the pulses in both the first and second control signal are a sub-harmonic frequency of the output signal.

In further embodiments, the invention minimizes DC offset voltages between the sampling modules during the differential sampling. In the serial configuration, this is accomplished by distributing a reference voltage to the input and output of the sampling modules. The result of minimizing (or preventing) DC offset voltages is that carrier insertion is minimized in the harmonics of the harmonically rich signal. In many transmit applications, carrier insertion is undesirable because the information to be transmitted is carried in the sidebands, and any energy at the carrier frequency is wasted. Alternatively, some transmit applications require sufficient carrier insertion for coherent demodulation of the transmitted signal at the receiver. In these applications, the invention can be configured to generate offset voltages between sampling modules, thereby causing carrier insertion in the harmonics of the harmonically rich signal.

An advantage is that embodiments of the invention up-convert a baseband signal directly from baseband-to-RF without any IF processing, while still meeting the spectral growth requirements of the most demanding communications standards. (Other embodiments may employ IF processing.) For example, in an I Q configuration, the invention can up-convert a CDMA spread spectrum signal directly from baseband-to-RF, and still meet the CDMA IS-95 figure-of-merit and spectral growth requirements. In other words, the invention is sufficiently linear and efficient during the up-conversion process that no IF filtering or amplification is required to meet the IS-95 figure-of-merit and spectral growth requirements. As a result, the entire IF chain in a conventional CDMA transmitter configuration can be eliminated, including the expensive and hard to integrate SAW filter. Since the SAW filter is eliminated, substantial portions of a CDMA transmitter that incorporate the invention can be integrated onto a single CMOS chip that uses a standard CMOS process, although the invention is not limited to this example application.

Further features and advantages of the invention, as well as the structure and operation of various embodiments of the invention, are described in detail below with reference to the accompanying drawings. The drawing in which an element first appears is typically indicated by the leftmost character(s) and/or digit(s) in the corresponding reference number.

#### BRIEF DESCRIPTION OF THE FIGURES

The present invention will be described with reference to the accompanying drawings, wherein:

FIG. 1A is a block diagram of a universal frequency translation (UFT) module according to an embodiment of the invention;

FIG. 1B is a more detailed diagram of a universal frequency translation (UFT) module according to an embodiment of the invention;

FIG. 1C illustrates a UFT module used in a universal frequency down-conversion (UFD) module according to an embodiment of the invention;

FIG. 1D illustrates a UFT module used in a universal frequency up-conversion (UFU) module according to an embodiment of the invention;

FIG. 2A is a block diagram of a universal frequency translation (UFT) module according to embodiments of the invention;

FIG. 2B is a block diagram of a universal frequency translation (UFT) module according to embodiments of the invention;

FIG. 3 is a block diagram of a universal frequency up-conversion (UFU) module according to an embodiment of the invention;

FIG. 4 is a more detailed diagram of a universal frequency up-conversion (UFU) module according to an embodiment of the invention;

FIG. 5 is a block diagram of a universal frequency up-conversion (UFU) module according to an alternative embodiment of the invention;

FIGS. 6A-6I illustrate example waveforms used to describe the operation of the UFU module;

FIG. 7 illustrates a UFT module used in a receiver according to an embodiment of the invention;

FIG. 8 illustrates a UFT module used in a transmitter according to an embodiment of the invention;

FIG. 9 illustrates an environment comprising a transmitter and a receiver, each of which may be implemented using a UFT module of the invention;

FIG. 10 illustrates a transceiver according to an embodiment of the invention;

FIG. 11 illustrates a transceiver according to an alternative embodiment of the invention;

FIG. 12 illustrates an environment comprising a transmitter and a receiver, each of which may be implemented using enhanced signal reception (ESR) components of the invention;

FIG. 13 illustrates a UFT module used in a unified down-conversion and filtering (UDF) module according to an embodiment of the invention;

FIG. 14 illustrates an example receiver implemented using a UDF module according to an embodiment of the invention;

FIGS. 15A-15F illustrate example applications of the UDF module according to embodiments of the invention;

FIG. 16 illustrates an environment comprising a transmitter and a receiver, each of which may be implemented using enhanced signal reception (ESR) components of the invention, wherein the receiver may be further implemented using one or more UFD modules of the invention;

FIG. 17 illustrates a unified down-converting and filtering (UDF) module according to an embodiment of the invention;

FIG. 18 is a table of example values at nodes in the UDF module of FIG. 17;

FIG. 19 is a detailed diagram of an example UDF module according to an embodiment of the invention;

FIGS. 20A and 20A-1 are example aliasing modules according to embodiments of the invention;

FIGS. 20B-20F are example waveforms used to describe the operation of the aliasing modules of FIGS. 20A and 20A-1;

FIG. 21 illustrates an enhanced signal reception system according to an embodiment of the invention;

FIGS. 22A-22F are example waveforms used to describe the system of FIG. 21;

FIG. 23A illustrates an example transmitter in an enhanced signal reception system according to an embodiment of the invention;

FIGS. 23B and 23C are example waveforms used to further describe the enhanced signal reception system according to an embodiment of the invention;

FIG. 23D illustrates another example transmitter in an enhanced signal reception system according to an embodiment of the invention;

FIGS. 23E and 23F are example waveforms used to further describe the enhanced signal reception system according to an embodiment of the invention;

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FIG. 24A illustrates an example receiver in an enhanced signal reception system according to an embodiment of the invention;

FIGS. 24B-24J are example waveforms used to further describe the enhanced signal reception system according to an embodiment of the invention;

FIGS. 25A-B illustrate carrier insertion;

FIGS. 26A-C illustrate a balanced transmitter 2602 according to an embodiment of the present invention;

FIG. 26B-C illustrate example waveforms that are associated with the balanced transmitter 2602 according to an embodiment of the present invention;

FIG. 26D illustrates example FET configurations of the balanced transmitter 2602;

FIGS. 27A-I illustrate various example timing diagrams associated with the transmitter 2602;

FIG. 27J illustrates an example frequency spectrum associated with the modulator 2604;

FIG. 28A illustrate a balanced modulator 2802 configured for carrier insertion according to embodiments of the present invention;

FIG. 28B illustrates example signal diagrams associated with the balanced transmitter 2802 according to embodiments of the invention;

FIG. 29 illustrates an I Q balanced transmitter 2920 according to embodiments of the present invention;

FIGS. 30A-C illustrate various example signal diagrams associated with the balanced transmitter 2920 in FIG. 29;

FIG. 31A illustrates an I Q balanced transmitter 3108 according to embodiments of the invention;

FIG. 31B illustrates an I Q balanced modulator 3118 according to embodiments of the invention;

FIG. 32 illustrates an I Q balanced modulator 3202 configured for carrier insertion according to embodiments of the invention;

FIG. 33 illustrates an I Q balanced modulator 3302 configured for carrier insertion according to embodiments of the invention;

FIGS. 34A-B illustrate various input configurations for the balanced transmitter 2920 according to embodiments of the present invention;

FIGS. 35A-B illustrate sidelobe requirements according to the IS-95 CDMA specification;

FIG. 36 illustrates a conventional CDMA transmitter 3600;

FIG. 37A illustrates a CDMA transmitter 3700 according to embodiments of the present invention;

FIGS. 37B-E illustrate various example signal diagrams according to embodiments of the present invention;

FIG. 37F illustrates a CDMA transmitter 3720 according to embodiments of the present invention;

FIG. 38 illustrates a CDMA transmitter utilizing a CMOS chip according to embodiments of the present invention;

FIG. 39 illustrates an example test set 3900;

FIGS. 40-52Z illustrate various example test results from testing the modulator 2910 in the test set 3900;

FIGS. 53A-C illustrate a transmitter 5300 and associated signal diagrams according to embodiments of the present invention;

FIGS. 54A-B illustrate a transmitter 5400 and associated signal diagrams according to embodiments of the present invention;

FIG. 54C illustrates a transmitter 5430 according to embodiments of the invention;

FIGS. 55A-D illustrates various implementation circuits for the modulator 2910 according to embodiments of the present invention;

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FIG. 56A illustrate a transmitter 5600 according to embodiments of the present invention;

FIGS. 56B-C illustrate various frequency spectrums that are associated with the transmitter 5600;

FIG. 56D illustrates a FET configuration for the modulator 5600;

FIG. 57 illustrates a IQ transmitter 5700 according to embodiments of the present invention;

FIGS. 58A-C illustrate various frequency spectrums that are associated with the IQ transmitter 5700;

FIG. 59 illustrates an IQ transmitter 5900 according to embodiments of the present invention;

FIG. 60 illustrates an IQ transmitter 6000 according to embodiments of the present invention;

FIG. 61 illustrates an IQ transmitter 6100 according to embodiments of the invention;

FIG. 62 illustrates a flowchart 6200 that is associated with the transmitter 2602 in the FIG. 26A according to an embodiment of the invention;

FIG. 63 illustrates a flowchart 6300 that further defines the flowchart 6200 in the FIG. 62, and is associated with the transmitter 2602 according to an embodiment of the invention;

FIG. 64 illustrates a flowchart 6400 that further defines the flowchart 6200 in the FIG. 63 and is associated with the transmitter 6400 according to an embodiment of the invention;

FIG. 65 illustrates the flowchart 6500 that is associated with the transmitter 2920 in the FIG. 29 according to an embodiment of the invention;

FIG. 66 illustrates a flowchart 6600 that is associated with the transmitter 5700 according to an embodiment of the invention;

FIG. 67 illustrates a flowchart 6700 that is associated with the spread spectrum transmitter 5300 in FIG. 53A according to an embodiment of the invention;

FIG. 68A and FIG. 68B illustrate a flowchart 6800 that is associated with an IQ spread spectrum modulator 6100 in FIG. 61 according to an embodiment of the invention;

FIG. 69A and FIG. 69B illustrate a flowchart 6900 that is associated with an IQ spread spectrum transmitter 5300 in FIG. 54A according to an embodiment of the invention;

FIG. 70A illustrates an IQ receiver having shunt UFT modules according to embodiments of the invention;

FIG. 70B illustrates control signal generator embodiments for receiver 7000 according to embodiments of the invention;

FIGS. 70C-D illustrate various control signal waveforms according to embodiments of the invention;

FIG. 70E illustrates an example IQ modulation receiver embodiment according to embodiments of the invention;

FIGS. 70E-P illustrate example waveforms that are representative of the IQ receiver in FIG. 70E;

FIGS. 70Q-R illustrate single channel receiver embodiments according to embodiments of the invention;

FIG. 71 illustrates a transceiver 7100 according to embodiments of the present invention;

FIG. 72 illustrates a transceiver 7200 according to embodiments of the present invention;

FIG. 73 illustrates a flowchart 7300 that is associated with the CDMA transmitter 3720 in FIG. 37 according to an embodiment of the invention;

FIG. 74A illustrates various pulse generators according to embodiments of the invention;

FIGS. 74B-C illustrate various example signal diagrams associated with the pulse generator in FIG. 74A, according to embodiments of the invention; and

FIGS. 74D-E illustrate various additional pulse generators according to embodiments of the invention.

## DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

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#### 1. Universal Frequency Translation

The present invention is related to frequency translation, and applications of same.

Such applications include, but are not limited to, frequency down-conversion, frequency up-conversion, enhanced signal reception, unified down-conversion and filtering, and combinations and applications of same.

FIG. 1A illustrates a universal frequency translation (UFT) module **102** according to embodiments of the invention. (The UFT module is also sometimes called a universal frequency translator, or a universal translator.)

As indicated by the example of FIG. 1A, some embodiments of the UFT module **102** include three ports (nodes), designated in FIG. 1A as Port **1**, Port **2**, and Port **3**. Other UFT embodiments include other than three ports.

Generally, the UFT module **102** (perhaps in combination with other components) operates to generate an output signal from an input signal, where the frequency of the output signal differs from the frequency of the input signal. In other words, the UFT module **102** (and perhaps other components) operates to generate the output signal from the input signal by translating the frequency (and perhaps other characteristics) of the input signal to the frequency (and perhaps other characteristics) of the output signal.

An example embodiment of the UFT module **103** is generally illustrated in FIG. 1B. Generally, the UFT module **103** includes a switch **106** controlled by a control signal **108**. The switch **106** is said to be a controlled switch.

As noted above, some UFT embodiments include other than three ports. For example, and without limitation, FIG. 2 illustrates an example UFT module **202**. The example UFT module **202** includes a diode **204** having two ports, designated as Port **1** and Port **2/3**. This embodiment does not include a third port, as indicated by the dotted line around the "Port **3**" label. FIG. 2B illustrates a second example UFT module **208** having a FET **210** whose gate is controlled by the control signal.

The UFT module is a very powerful and flexible device. Its flexibility is illustrated, in part, by the wide range of applications in which it can be used. Its power is illustrated, in part, by the usefulness and performance of such applications.

For example, a UFT module **115** can be used in a universal frequency down-conversion (UFD) module **114**, an example of which is shown in FIG. 1C. In this capacity, the UFT module **115** frequency down-converts an input signal to an output signal.

As another example, as shown in FIG. 1D, a UFT module **117** can be used in a universal frequency up-conversion (UFU) module **116**. In this capacity, the UFT module **117** frequency up-converts an input signal to an output signal.

These and other applications of the UFT module are described below. Additional applications of the UFT module will be apparent to persons skilled in the relevant art(s) based on the teachings contained herein. In some applications, the UFT module is a required component. In other applications, the UFT module is an optional component.

#### 2. Frequency Down-conversion

The present invention is directed to systems and methods of universal frequency down-conversion, and applications of same.

In particular, the following discussion describes down-converting using a Universal Frequency Translation Module. The down-conversion of an EM signal by aliasing the EM signal at an aliasing rate is fully described in co-pending U.S. patent application entitled "Method and System for Down-Converting Electromagnetic Signals," Ser. No. 09/176,022, filed Oct. 21, 1998, the full disclosure of which is incorporated herein by reference. A relevant portion of the above mentioned patent application is summarized below to describe down-converting an input signal to produce a down-converted signal that exists at a lower frequency or a baseband signal.

FIG. 20A illustrates an aliasing module **2000** (one embodiment of a UFD module) for down-conversion using a universal frequency translation (UFT) module **2002**, which down-converts an EM input signal **2004**. In particular embodiments, aliasing module **2000** includes a switch **2008** and a capacitor **2010**. The electronic alignment of the circuit components is flexible. That is, in one implementation, the switch **2008** is in series with input signal **2004** and capacitor **2010** is shunted to ground (although it may be other than ground in configurations such as differential mode). In a second implementation

(see FIG. 20A-1), the capacitor 2010 is in series with the input signal 2004 and the switch 2008 is shunted to ground (although it may be other than ground in configurations such as differential mode). Aliasing module 2000 with UFT module 2002 can be easily tailored to down-convert a wide variety of electromagnetic signals using aliasing frequencies that are well below the frequencies of the EM input signal 2004.

In one implementation, aliasing module 2000 down-converts the input signal 2004 to an intermediate frequency (IF) signal. In another implementation, the aliasing module 2000 down-converts the input signal 2004 to a demodulated baseband signal. In yet another implementation, the input signal 2004 is a frequency modulated (FM) signal, and the aliasing module 2000 down-converts it to a non-FM signal, such as a phase modulated (PM) signal or an amplitude modulated (AM) signal. Each of the above implementations is described below.

In an embodiment, the control signal 2006 includes a train of pulses that repeat at an aliasing rate that is equal to, or less than, twice the frequency of the input signal 2004. In this embodiment, the control signal 2006 is referred to herein as an aliasing signal because it is below the Nyquist rate for the frequency of the input signal 2004. Preferably, the frequency of control signal 2006 is much less than the input signal 2004.

A train of pulses 2018 as shown in FIG. 20D controls the switch 2008 to alias the input signal 2004 with the control signal 2006 to generate a down-converted output signal 2012. More specifically, in an embodiment, switch 2008 closes on a first edge of each pulse 2020 of FIG. 20D and opens on a second edge of each pulse. When the switch 2008 is closed, the input signal 2004 is coupled to the capacitor 2010, and charge is transferred from the input signal to the capacitor 2010. The charge stored during successive pulses forms down-converted output signal 2012.

Exemplary waveforms are shown in FIGS. 20B-20F.

FIG. 20B illustrates an analog amplitude modulated (AM) carrier signal 2014 that is an example of input signal 2004. For illustrative purposes, in FIG. 20C, an analog AM carrier signal portion 2016 illustrates a portion of the analog AM carrier signal 2014 on an expanded time scale. The analog AM carrier signal portion 2016 illustrates the analog AM carrier signal 2014 from time  $t_0$  to time  $t_1$ .

FIG. 20D illustrates an exemplary aliasing signal 2018 that is an example of control signal 2006. Aliasing signal 2018 is on approximately the same time scale as the analog AM carrier signal portion 2016. In the example shown in FIG. 20D, the aliasing signal 2018 includes a train of pulses 2020 having negligible apertures that tend towards zero (the invention is not limited to this embodiment, as discussed below). The pulse aperture may also be referred to as the pulse width as will be understood by those skilled in the art(s). The pulses 2020 repeat at an aliasing rate, or pulse repetition rate of aliasing signal 2018. The aliasing rate is determined as described below, and further described in co-pending U.S. patent application entitled "Method and System for Down-converting Electromagnetic Signals," application Ser. No. 09/176,022.

As noted above, the train of pulses 2020 (i.e., control signal 2006) control the switch 2008 to alias the analog AM carrier signal 2016 (i.e., input signal 2004) at the aliasing rate of the aliasing signal 2018. Specifically, in this embodiment, the switch 2008 closes on a first edge of each pulse and opens on a second edge of each pulse. When the switch 2008 is closed, input signal 2004 is coupled to the capacitor 2010, and charge is transferred from the input signal 2004 to the capacitor 2010. The charge transferred during a pulse is referred to herein as an under-sample. Exemplary under-samples 2022

form down-converted signal portion 2024 (FIG. 20E) that corresponds to the analog AM carrier signal portion 2016 (FIG. 20C) and the train of pulses 2020 (FIG. 20D). The charge stored during successive under-samples of AM carrier signal 2014 form the down-converted signal 2024 (FIG. 20E) that is an example of down-converted output signal 2012 (FIG. 20A). In FIG. 20F, a demodulated baseband signal 2026 represents the demodulated baseband signal 2024 after filtering on a compressed time scale. As illustrated, down-converted signal 2026 has substantially the same "amplitude envelope" as AM carrier signal 2014. Therefore, FIGS. 20B-20F illustrate down-conversion of AM carrier signal 2014.

The waveforms shown in FIGS. 20B-20F are discussed herein for illustrative purposes only, and are not limiting. Additional exemplary time domain and frequency domain drawings, and exemplary methods and systems of the invention relating thereto, are disclosed in co-pending U.S. patent application entitled "Method and System for Down-converting Electromagnetic Signals," application Ser. No. 09/176,022.

The aliasing rate of control signal 2006 determines whether the input signal 2004 is down-converted to an IF signal, down-converted to a demodulated baseband signal, or down-converted from an FM signal to a PM or an AM signal. Generally, relationships between the input signal 2004, the aliasing rate of the control signal 2006, and the down-converted output signal 2012 are illustrated below:

$$(\text{Freq. of input signal } 2004) = n \cdot (\text{Freq. of control signal } 2006) \pm (\text{Freq. of Down-Converted Output Signal } 2012)$$

For the examples contained herein, only the "+" condition will be discussed. The value of  $n$  represents a harmonic or sub-harmonic of input signal 2004 (e.g.,  $n=0.5, 1, 2, 3, \dots$ ).

When the aliasing rate of control signal 2006 is off-set from the frequency of input signal 2004, or off-set from a harmonic or sub-harmonic thereof, input signal 2004 is down-converted to an IF signal. This is because the under-sampling pulses occur at different phases of subsequent cycles of input signal 2004. As a result, the under-samples form a lower frequency oscillating pattern. If the input signal 2004 includes lower frequency changes, such as amplitude, frequency, phase, etc., or any combination thereof, the charge stored during associated under-samples reflects the lower frequency changes, resulting in similar changes on the down-converted IF signal. For example, to down-convert a 901 MHz input signal to a 1 MHz IF signal, the frequency of the control signal 2006 would be calculated as follows:

$$(\text{Freq}_{\text{input}} - \text{Freq}_{\text{IF}}) / n = \text{Freq}_{\text{control}}$$

$$(901 \text{ MHz} - 1 \text{ MHz}) / n = 900 / n$$

For  $n=0.5, 1, 2, 3, 4, \dots$ , the frequency of the control signal 2006 would be substantially equal to 1.8 GHz, 900 MHz, 450 MHz, 300 MHz, 225 MHz, etc.

Exemplary time domain and frequency domain drawings, illustrating down-conversion of analog and digital AM, PM and FM signals to IF signals, and exemplary methods and systems thereof, are disclosed in co-pending U.S. patent application entitled "Method and System for Down-converting Electromagnetic Signals," application Ser. No. 09/176,022.

Alternatively, when the aliasing rate of the control signal 2006 is substantially equal to the frequency of the input signal 2004, or substantially equal to a harmonic or sub-harmonic thereof, input signal 2004 is directly down-converted to a demodulated baseband signal. This is because, without modulation, the under-sampling pulses occur at the same

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point of subsequent cycles of the input signal **2004**. As a result, the under-samples form a constant output baseband signal. If the input signal **2004** includes lower frequency changes, such as amplitude, frequency, phase, etc., or any combination thereof, the charge stored during associated under-samples reflects the lower frequency changes, resulting in similar changes on the demodulated baseband signal. For example, to directly down-convert a 900 MHz input signal to a demodulated baseband signal (i.e., zero IF), the frequency of the control signal **2006** would be calculated as follows:

$$(\text{Freq}_{\text{input}} - \text{Freq}_{\text{IF}}) / n = \text{Freq}_{\text{control}}$$

$$(900 \text{ MHz} - 0 \text{ MHz}) / n = 900 \text{ MHz} / n$$

For  $n=0.5, 1, 2, 3, 4$ , etc., the frequency of the control signal **2006** should be substantially equal to 1.8 GHz, 900 MHz, 450 MHz, 300 MHz, 225 MHz, etc.

Exemplary time domain and frequency domain drawings, illustrating direct down-conversion of analog and digital AM and PM signals to demodulated baseband signals, and exemplary methods and systems thereof, are disclosed in the co-pending U.S. patent application entitled "Method and System for Down-converting Electromagnetic Signals," application Ser. No. 09/176,022.

Alternatively, to down-convert an input FM signal to a non-FM signal, a frequency within the FM bandwidth must be down-converted to baseband (i.e., zero IF). As an example, to down-convert a frequency shift keying (FSK) signal (a sub-set of FM) to a phase shift keying (PSK) signal (a subset of PM), the mid-point between a lower frequency  $F_1$  and an upper frequency  $F_2$  (that is,  $[(F_1 + F_2) / 2]$ ) of the FSK signal is down-converted to zero IF. For example, to down-convert an FSK signal having  $F_1$  equal to 899 MHz and  $F_2$  equal to 901 MHz, to a PSK signal, the aliasing rate of the control signal **2006** would be calculated as follows:

$$\begin{aligned} \text{Frequency of the input} &= (F_1 + F_2) \div 2 \\ &= (899 \text{ MHz} + 901 \text{ MHz}) \div 2 \\ &= 900 \text{ MHz} \end{aligned}$$

Frequency of the down-converted signal=0 (i.e., baseband)

$$(\text{Freq}_{\text{input}} - \text{Freq}_{\text{IF}}) / n = \text{Freq}_{\text{control}}$$

$$(900 \text{ MHz} - 0 \text{ MHz}) / n = 900 \text{ MHz} / n$$

For  $n=0.5, 1, 2, 3$ , etc., the frequency of the control signal **2006** should be substantially equal to 1.8 GHz, 900 MHz, 450 MHz, 300 MHz, 225 MHz, etc. The frequency of the down-converted PSK signal is substantially equal to one half the difference between the lower frequency  $F_1$  and the upper frequency  $F_2$ .

As another example, to down-convert a FSK signal to an amplitude shift keying (ASK) signal (a subset of AM), either the lower frequency  $F_1$  or the upper frequency  $F_2$  of the FSK signal is down-converted to zero IF. For example, to down-convert an FSK signal having  $F_1$  equal to 900 MHz and  $F_2$  equal to 901 MHz, to an ASK signal, the aliasing rate of the control signal **2006** should be substantially equal to:

$$(900 \text{ MHz} - 0 \text{ MHz}) / n = 900 \text{ MHz} / n, \text{ or}$$

$$(901 \text{ MHz} - 0 \text{ MHz}) / n = 901 \text{ MHz} / n.$$

For the former case of 900 MHz/n, and for  $n=0.5, 1, 2, 3, 4$ , etc., the frequency of the control signal **2006** should be sub-

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stantially equal to 1.8 GHz, 900 MHz, 450 MHz, 300 MHz, 225 MHz, etc. For the latter case of 901 MHz/n, and for  $n=0.5, 1, 2, 3, 4$ , etc., the frequency of the control signal **2006** should be substantially equal to 1.802 GHz, 901 MHz, 450.5 MHz, 300.333 MHz, 225.25 MHz, etc. The frequency of the down-converted AM signal is substantially equal to the difference between the lower frequency  $F_1$  and the upper frequency  $F_2$  (i.e., 1 MHz).

Exemplary time domain and frequency domain drawings, illustrating down-conversion of FM signals to non-FM signals, and exemplary methods and systems thereof, are disclosed in the co-pending U.S. patent application entitled "Method and System for Down-converting Electromagnetic Signals," application Ser. No. 09/176,022.

In an embodiment, the pulses of the control signal **2006** have negligible apertures that tend towards zero. This makes the UFT module **2002** a high input impedance device. This configuration is useful for situations where minimal disturbance of the input signal may be desired.

In another embodiment, the pulses of the control signal **2006** have non-negligible apertures that tend away from zero. This makes the UFT module **2002** a lower input impedance device. This allows the lower input impedance of the UFT module **2002** to be substantially matched with a source impedance of the input signal **2004**. This also improves the energy transfer from the input signal **2004** to the down-converted output signal **2012**, and hence the efficiency and signal to noise (s/n) ratio of UFT module **2002**.

Exemplary systems and methods for generating and optimizing the control signal **2006**, and for otherwise improving energy transfer and s/n ratio, are disclosed in the co-pending U.S. patent application entitled "Method and System for Down-converting Electromagnetic Signals," application Ser. No. 09/176,022.

### 3. Frequency Up-conversion Using Universal Frequency Translation

The present invention is directed to systems and methods of frequency up-conversion, and applications of same.

An example frequency up-conversion system **300** is illustrated in FIG. 3. The frequency up-conversion system **300** is now described.

An input signal **302** (designated as "Control Signal" in FIG. 3) is accepted by a switch module **304**. For purposes of example only, assume that the input signal **302** is a FM input signal **606**, an example of which is shown in FIG. 6C. FM input signal **606** may have been generated by modulating information signal **602** onto oscillating signal **604** (FIGS. 6A and 6B). It should be understood that the invention is not limited to this embodiment. The information signal **602** can be analog, digital, or any combination thereof, and any modulation scheme can be used.

The output of switch module **304** is a harmonically rich signal **306**, shown for example in FIG. 6D as a harmonically rich signal **608**. The harmonically rich signal **608** has a continuous and periodic waveform.

FIG. 6E is an expanded view of two sections of harmonically rich signal **608**, section **610** and section **612**. The harmonically rich signal **608** may be a rectangular wave, such as a square wave or a pulse (although, the invention is not limited to this embodiment). For ease of discussion, the term "rectangular waveform" is used to refer to waveforms that are substantially rectangular. In a similar manner, the term "square wave" refers to those waveforms that are substantially square and it is not the intent of the present invention that a perfect square wave be generated or needed.

Harmonically rich signal **608** is comprised of a plurality of sinusoidal waves whose frequencies are integer multiples of

the fundamental frequency of the waveform of the harmonically rich signal 608. These sinusoidal waves are referred to as the harmonics of the underlying waveform, and the fundamental frequency is referred to as the first harmonic. FIG. 6F and FIG. 6G show separately the sinusoidal components making up the first, third, and fifth harmonics of section 610 and section 612. (Note that in theory there may be an infinite number of harmonics; in this example, because harmonically rich signal 608 is shown as a square wave, there are only odd harmonics). Three harmonics are shown simultaneously (but not summed) in FIG. 6H.

The relative amplitudes of the harmonics are generally a function of the relative widths of the pulses of harmonically rich signal 306 and the period of the fundamental frequency, and can be determined by doing a Fourier analysis of harmonically rich signal 306. According to an embodiment of the invention, the input signal 606 may be shaped to ensure that the amplitude of the desired harmonic is sufficient for its intended use (e.g., transmission).

A filter 308 filters out any undesired frequencies (harmonics), and outputs an electromagnetic (EM) signal at the desired harmonic frequency or frequencies as an output signal 310, shown for example as a filtered output signal 614 in FIG. 6I.

FIG. 4 illustrates an example universal frequency up-conversion (UFU) module 401. The UFU module 401 includes an example switch module 304, which comprises a bias signal 402, a resistor or impedance 404, a universal frequency translator (UFT) 450, and a ground 408. The UFT 450 includes a switch 406. The input signal 302 (designated as "Control Signal" in FIG. 4) controls the switch 406 in the UFT 450, and causes it to close and open. Harmonically rich signal 306 is generated at a node 405 located between the resistor or impedance 404 and the switch 406.

Also in FIG. 4, it can be seen that an example filter 308 is comprised of a capacitor 410 and an inductor 412 shunted to a ground 414. The filter is designed to filter out the undesired harmonics of harmonically rich signal 306.

The invention is not limited to the UFU embodiment shown in FIG. 4.

For example, in an alternate embodiment shown in FIG. 5, an unshaped input signal 501 is routed to a pulse shaping module 502. The pulse shaping module 502 modifies the unshaped input signal 501 to generate a (modified) input signal 302 (designated as the "Control Signal" in FIG. 5). The input signal 302 is routed to the switch module 304, which operates in the manner described above. Also, the filter 308 of FIG. 5 operates in the manner described above.

The purpose of the pulse shaping module 502 is to define the pulse width of the input signal 302. Recall that the input signal 302 controls the opening and closing of the switch 406 in switch module 304. During such operation, the pulse width of the input signal 302 establishes the pulse width of the harmonically rich signal 306. As stated above, the relative amplitudes of the harmonics of the harmonically rich signal 306 are a function of at least the pulse width of the harmonically rich signal 306. As such, the pulse width of the input signal 302 contributes to setting the relative amplitudes of the harmonics of harmonically rich signal 306.

Further details of up-conversion as described in this section are presented in pending U.S. application "Method and System for Frequency Up-Conversion," Ser. No. 09/176,154, filed Oct. 21, 1998, incorporated herein by reference in its entirety.

#### 4. Enhanced Signal Reception

The present invention is directed to systems and methods of enhanced signal reception (ESR), and applications of same.

Referring to FIG. 21, transmitter 2104 accepts a modulating baseband signal 2102 and generates (transmitted) redundant spectrums 2106a-n, which are sent over communications medium 2108. Receiver 2112 recovers a demodulated baseband signal 2114 from (received) redundant spectrums 2110a-n. Demodulated baseband signal 2114 is representative of the modulating baseband signal 2102, where the level of similarity between the modulating baseband signal 2114 and the modulating baseband signal 2102 is application dependent.

Modulating baseband signal 2102 is preferably any information signal desired for transmission and/or reception. An example modulating baseband signal 2202 is illustrated in FIG. 22A, and has an associated modulating baseband spectrum 2204 and image spectrum 2203 that are illustrated in FIG. 22B. Modulating baseband signal 2202 is illustrated as an analog signal in FIG. 22a, but could also be a digital signal, or combination thereof. Modulating baseband signal 2202 could be a voltage (or current) characterization of any number of real world occurrences, including for example and without limitation, the voltage (or current) representation for a voice signal.

Each transmitted redundant spectrum 2106a-n contains the necessary information to substantially reconstruct the modulating baseband signal 2102. In other words, each redundant spectrum 2106a-n contains the necessary amplitude, phase, and frequency information to reconstruct the modulating baseband signal 2102.

FIG. 22C illustrates example transmitted redundant spectrums 2206b-d. Transmitted redundant spectrums 2206b-d are illustrated to contain three redundant spectrums for illustration purposes only. Any number of redundant spectrums could be generated and transmitted as will be explained in following discussions.

Transmitted redundant spectrums 2206b-d are centered at  $f_1$ , with a frequency spacing  $f_2$  between adjacent spectrums. Frequencies  $f_1$  and  $f_2$  are dynamically adjustable in real-time as will be shown below. FIG. 22D illustrates an alternate embodiment, where redundant spectrums 2208c,d are centered on unmodulated oscillating signal 2209 at  $f_1$  (Hz). Oscillating signal 2209 may be suppressed if desired using, for example, phasing techniques or filtering techniques. Transmitted redundant spectrums are preferably above baseband frequencies as is represented by break 2205 in the frequency axis of FIGS. 22C and 22D.

Received redundant spectrums 2110a-n are substantially similar to transmitted redundant spectrums 2106a-n, except for the changes introduced by the communications medium 2108. Such changes can include but are not limited to signal attenuation, and signal interference. FIG. 22E illustrates example received redundant spectrums 2210b-d. Received redundant spectrums 2210b-d are substantially similar to transmitted redundant spectrums 2206b-d, except that redundant spectrum 2210c includes an undesired jamming signal spectrum 2211 in order to illustrate some advantages of the present invention. Jamming signal spectrum 2211 is a frequency spectrum associated with a jamming signal. For purposes of this invention, a "jamming signal" refers to any unwanted signal, regardless of origin, that may interfere with the proper reception and reconstruction of an intended signal. Furthermore, the jamming signal is not limited to tones as depicted by spectrum 2211, and can have any spectral shape, as will be understood by those skilled in the art(s).

As stated above, demodulated baseband signal **2114** is extracted from one or more of received redundant spectrums **2210b-d**. FIG. **22F** illustrates example demodulated baseband signal **2212** that is, in this example, substantially similar to modulating baseband signal **2202** (FIG. **22A**); where in practice, the degree of similarity is application dependent.

An advantage of the present invention should now be apparent. The recovery of modulating baseband signal **2202** can be accomplished by receiver **2112** in spite of the fact that high strength jamming signal(s) (e.g. jamming signal spectrum **2211**) exist on the communications medium. The intended baseband signal can be recovered because multiple redundant spectrums are transmitted, where each redundant spectrum carries the necessary information to reconstruct the baseband signal. At the destination, the redundant spectrums are isolated from each other so that the baseband signal can be recovered even if one or more of the redundant spectrums are corrupted by a jamming signal.

Transmitter **2104** will now be explored in greater detail. FIG. **23A** illustrates transmitter **2301**, which is one embodiment of transmitter **2104** that generates redundant spectrums configured similar to redundant spectrums **2206b-d**. Transmitter **2301** includes generator **2303**, optional spectrum processing module **2304**, and optional medium interface module **2320**. Generator **2303** includes: first oscillator **2302**, second oscillator **2309**, first stage modulator **2306**, and second stage modulator **2310**.

Transmitter **2301** operates as follows. First oscillator **2302** and second oscillator **2309** generate a first oscillating signal **2305** and second oscillating signal **2312**, respectively. First stage modulator **2306** modulates first oscillating signal **2305** with modulating baseband signal **2202**, resulting in modulated signal **2308**. First stage modulator **2306** may implement any type of modulation including but not limited to: amplitude modulation, frequency modulation, phase modulation, combinations thereof, or any other type of modulation. Second stage modulator **2310** modulates modulated signal **2308** with second oscillating signal **2312**, resulting in multiple redundant spectrums **2206a-n** shown in FIG. **23B**. Second stage modulator **2310** is preferably a phase modulator, or a frequency modulator, although other types of modulation may be implemented including but not limited to amplitude modulation. Each redundant spectrum **2206a-n** contains the necessary amplitude, phase, and frequency information to substantially reconstruct the modulating baseband signal **2202**.

Redundant spectrums **2206a-n** are substantially centered around  $f_1$ , which is the characteristic frequency of first oscillating signal **2305**. Also, each redundant spectrum **2206a-n** (except for **2206c**) is offset from  $f_1$  by approximately a multiple of  $f_2$  (Hz), where  $f_2$  is the frequency of the second oscillating signal **2312**. Thus, each redundant spectrum **2206a-n** is offset from an adjacent redundant spectrum by  $f_2$  (Hz). This allows the spacing between adjacent redundant spectrums to be adjusted (or tuned) by changing  $f_2$  that is associated with second oscillator **2309**. Adjusting the spacing between adjacent redundant spectrums allows for dynamic real-time tuning of the bandwidth occupied by redundant spectrums **2206a-n**.

In one embodiment, the number of redundant spectrums **2206a-n** generated by transmitter **2301** is arbitrary and may be unlimited as indicated by the "a-n" designation for redundant spectrums **2206a-n**. However, a typical communications medium will have a physical and/or administrative limitations (i.e. FCC regulations) that restrict the number of redundant spectrums that can be practically transmitted over the communications medium. Also, there may be other reasons to

limit the number of redundant spectrums transmitted. Therefore, preferably, the transmitter **2301** will include an optional spectrum processing module **2304** to process the redundant spectrums **2206a-n** prior to transmission over communications medium **2108**.

In one embodiment, spectrum processing module **2304** includes a filter with a passband **2207** (FIG. **23C**) to select redundant spectrums **2206b-d** for transmission. This will substantially limit the frequency bandwidth occupied by the redundant spectrums to the passband **2207**. In one embodiment, spectrum processing module **2304** also up converts redundant spectrums and/or amplifies redundant spectrums prior to transmission over the communications medium **2108**. Finally, medium interface module **2320** transmits redundant spectrums over the communications medium **2108**. In one embodiment, communications medium **2108** is an over-the-air link and medium interface module **2320** is an antenna. Other embodiments for communications medium **2108** and medium interface module **2320** will be understood based on the teachings contained herein.

FIG. **23D** illustrates transmitter **2321**, which is one embodiment of transmitter **2104** that generates redundant spectrums configured similar to redundant spectrums **2208c-d** and unmodulated spectrum **2209**. Transmitter **2321** includes generator **2311**, spectrum processing module **2304**, and (optional) medium interface module **2320**. Generator **2311** includes: first oscillator **2302**, second oscillator **2309**, first stage modulator **2306**, and second stage modulator **2310**.

As shown in FIG. **23D**, many of the components in transmitter **2321** are similar to those in transmitter **2301**. However, in this embodiment, modulating baseband signal **2202** modulates second oscillating signal **2312**. Transmitter **2321** operates as follows. First stage modulator **2306** modulates second oscillating signal **2312** with modulating baseband signal **2202**, resulting in modulated signal **2322**. As described earlier, first stage modulator **2306** can effect any type of modulation including but not limited to: amplitude modulation, frequency modulation, combinations thereof, or any other type of modulation. Second stage modulator **2310** modulates first oscillating signal **2304** with modulated signal **2322**, resulting in redundant spectrums **2208a-n**, as shown in FIG. **23E**. Second stage modulator **2310** is preferably a phase or frequency modulator, although other modulators could be used including but not limited to an amplitude modulator.

Redundant spectrums **2208a-n** are centered on unmodulated spectrum **2209** (at  $f_1$  Hz), and adjacent spectrums are separated by  $f_2$  Hz. The number of redundant spectrums **2208a-n** generated by generator **2311** is arbitrary and unlimited, similar to spectrums **2206a-n** discussed above. Therefore, optional spectrum processing module **2304** may also include a filter with passband **2325** to select, for example, spectrums **2208c,d** for transmission over communications medium **2108**. In addition, optional spectrum processing module **2304** may also include a filter (such as a bandstop filter) to attenuate unmodulated spectrum **2209**. Alternatively, unmodulated spectrum **2209** may be attenuated by using phasing techniques during redundant spectrum generation. Finally, (optional) medium interface module **2320** transmits redundant spectrums **2208c,d** over communications medium **2108**.

Receiver **2112** will now be explored in greater detail to illustrate recovery of a demodulated baseband signal from received redundant spectrums. FIG. **24A** illustrates receiver **2430**, which is one embodiment of receiver **2112**. Receiver **2430** includes optional medium interface module **2402**, down-converter **2404**, spectrum isolation module **2408**, and data extraction module **2414**. Spectrum isolation module

**2408** includes filters **2410a-c**. Data extraction module **2414** includes demodulators **2416a-c**, error check modules **2420a-c**, and arbitration module **2424**. Receiver **2430** will be discussed in relation to the signal diagrams in FIGS. **24B-24J**.

In one embodiment, optional medium interface module **2402** receives redundant spectrums **2210b-d** (FIG. **22E**, and FIG. **24B**). Each redundant spectrum **2210b-d** includes the necessary amplitude, phase, and frequency information to substantially reconstruct the modulating baseband signal used to generate the redundant spectrums. However, in the present example, spectrum **2210c** also contains jamming signal **2211**, which may interfere with the recovery of a baseband signal from spectrum **2210c**. Down-converter **2404** down-converts received redundant spectrums **2210b-d** to lower intermediate frequencies, resulting in redundant spectrums **2406a-c** (FIG. **24C**). Jamming signal **2211** is also down-converted to jamming signal **2407**, as it is contained within redundant spectrum **2406b**. Spectrum isolation module **2408** includes filters **2410a-c** that isolate redundant spectrums **2406a-c** from each other (FIGS. **24D-24F**, respectively). Demodulators **2416a-c** independently demodulate spectrums **2406a-c**, resulting in demodulated baseband signals **2418a-c**, respectively (FIGS. **24G-24I**). Error check modules **2420a-c** analyze demodulated baseband signal **2418a-c** to detect any errors. In one embodiment, each error check module **2420a-c** sets an error flag **2422a-c** whenever an error is detected in a demodulated baseband signal. Arbitration module **2424** accepts the demodulated baseband signals and associated error flags, and selects a substantially error-free demodulated baseband signal (FIG. **24J**). In one embodiment, the substantially error-free demodulated baseband signal will be substantially similar to the modulating baseband signal used to generate the received redundant spectrums, where the degree of similarity is application dependent.

Referring to FIGS. **24G-I**, arbitration module **2424** will select either demodulated baseband signal **2418a** or **2418c**, because error check module **2420b** will set the error flag **2422b** that is associated with demodulated baseband signal **2418b**.

The error detection schemes implemented by the error detection modules include but are not limited to: cyclic redundancy check (CRC) and parity check for digital signals, and various error detection schemes for analog signal.

Further details of enhanced signal reception as described in this section are presented in pending U.S. application "Method and System for Ensuring Reception of a Communications Signal," Ser. No. 09/176,415, filed Oct. 21, 1998, incorporated herein by reference in its entirety.

#### 5. Unified Down-conversion And Filtering

The present invention is directed to systems and methods of unified down-conversion and filtering (UDF), and applications of same.

In particular, the present invention includes a unified down-converting and filtering (UDF) module that performs frequency selectivity and frequency translation in a unified (i.e., integrated) manner. By operating in this manner, the invention achieves high frequency selectivity prior to frequency translation (the invention is not limited to this embodiment). The invention achieves high frequency selectivity at substantially any frequency, including but not limited to RF (radio frequency) and greater frequencies. It should be understood that the invention is not limited to this example of RF and greater frequencies. The invention is intended, adapted, and capable of working with lower than radio frequencies.

FIG. **17** is a conceptual block diagram of a UDF module **1702** according to an embodiment of the present invention.

The UDF module **1702** performs at least frequency translation and frequency selectivity.

The effect achieved by the UDF module **1702** is to perform the frequency selectivity operation prior to the performance of the frequency translation operation. Thus, the UDF module **1702** effectively performs input filtering.

According to embodiments of the present invention, such input filtering involves a relatively narrow bandwidth. For example, such input filtering may represent channel select filtering, where the filter bandwidth may be, for example, 50 KHz to 150 KHz. It should be understood, however, that the invention is not limited to these frequencies. The invention is intended, adapted, and capable of achieving filter bandwidths of less than and greater than these values.

In embodiments of the invention, input signals **1704** received by the UDF module **1702** are at radio frequencies. The UDF module **1702** effectively operates to input filter these RF input signals **1704**. Specifically, in these embodiments, the UDF module **1702** effectively performs input, channel select filtering of the RF input signal **1704**. Accordingly, the invention achieves high selectivity at high frequencies.

The UDF module **1702** effectively performs various types of filtering, including but not limited to bandpass filtering, low pass filtering, high pass filtering, notch filtering, all pass filtering, band stop filtering, etc., and combinations thereof.

Conceptually, the UDF module **1702** includes a frequency translator **1708**. The frequency translator **1708** conceptually represents that portion of the UDF module **1702** that performs frequency translation (down conversion).

The UDF module **1702** also conceptually includes an apparent input filter **1706** (also sometimes called an input filtering emulator). Conceptually, the apparent input filter **1706** represents that portion of the UDF module **1702** that performs input filtering.

In practice, the input filtering operation performed by the UDF module **1702** is integrated with the frequency translation operation. The input filtering operation can be viewed as being performed concurrently with the frequency translation operation. This is a reason why the input filter **1706** is herein referred to as an "apparent" input filter **1706**.

The UDF module **1702** of the present invention includes a number of advantages. For example, high selectivity at high frequencies is realizable using the UDF module **1702**. This feature of the invention is evident by the high Q factors that are attainable. For example, and without limitation, the UDF module **1702** can be designed with a filter center frequency  $f_c$  on the order of 900 MHz, and a filter bandwidth on the order of 50 KHz. This represents a Q of 18,000 (Q is equal to the center frequency divided by the bandwidth).

It should be understood that the invention is not limited to filters with high Q factors. The filters contemplated by the present invention may have lesser or greater Qs, depending on the application, design, and/or implementation. Also, the scope of the invention includes filters where Q factor as discussed herein is not applicable.

The invention exhibits additional advantages. For example, the filtering center frequency  $f_c$  of the UDF module **1702** can be electrically adjusted, either statically or dynamically.

Also, the UDF module **1702** can be designed to amplify input signals.

Further, the UDF module **1702** can be implemented without large resistors, capacitors, or inductors. Also, the UDF module **1702** does not require that tight tolerances be maintained on the values of its individual components, i.e., its resistors, capacitors, inductors, etc. As a result, the architec-

ture of the UDF module **1702** is friendly to integrated circuit design techniques and processes.

The features and advantages exhibited by the UDF module **1702** are achieved at least in part by adopting a new technological paradigm with respect to frequency selectivity and translation. Specifically, according to the present invention, the UDF module **1702** performs the frequency selectivity operation and the frequency translation operation as a single, unified (integrated) operation. According to the invention, operations relating to frequency translation also contribute to the performance of frequency selectivity, and vice versa.

According to embodiments of the present invention, the UDF module generates an output signal from an input signal using samples/instances of the input signal and samples/instances of the output signal.

More particularly, first, the input signal is under-sampled. This input sample includes information (such as amplitude, phase, etc.) representative of the input signal existing at the time the sample was taken.

As described further below, the effect of repetitively performing this step is to translate the frequency (that is, down-convert) of the input signal to a desired lower frequency, such as an intermediate frequency (IF) or baseband.

Next, the input sample is held (that is, delayed).

Then, one or more delayed input samples (some of which may have been scaled) are combined with one or more delayed instances of the output signal (some of which may have been scaled) to generate a current instance of the output signal.

Thus, according to a preferred embodiment of the invention, the output signal is generated from prior samples/instances of the input signal and/or the output signal. (It is noted that, in some embodiments of the invention, current samples/instances of the input signal and/or the output signal may be used to generate current instances of the output signal.) By operating in this manner, the UDF module preferably performs input filtering and frequency down-conversion in a unified manner.

FIG. **19** illustrates an example implementation of the unified down-converting and filtering (UDF) module **1922**. The UDF module **1922** performs the frequency translation operation and the frequency selectivity operation in an integrated, unified manner as described above, and as further described below.

In the example of FIG. **19**, the frequency selectivity operation performed by the UDF module **1922** comprises a band-pass filtering operation according to EQ. 1, below, which is an example representation of a band-pass filtering transfer function.

$$VO = \alpha_1 z^{-1} VI - \beta_1 z^{-1} VO - \beta_0 z^{-2} VO \quad \text{EQ. 1}$$

It should be noted, however, that the invention is not limited to band-pass filtering.

Instead, the invention effectively performs various types of filtering, including but not limited to bandpass filtering, low pass filtering, high pass filtering, notch filtering, all pass filtering, band stop filtering, etc., and combinations thereof. As will be appreciated, there are many representations of any given filter type. The invention is applicable to these filter representations. Thus, EQ. 1 is referred to herein for illustrative purposes only, and is not limiting.

The UDF module **1922** includes a down-convert and delay module **1924**, first and second delay modules **1928** and **1930**, first and second scaling modules **1932** and **1934**, an output sample and hold module **1936**, and an (optional) output smoothing module **1938**. Other embodiments of the UDF module will have these components in different configura-

tions, and/or a subset of these components, and/or additional components. For example, and without limitation, in the configuration shown in FIG. **19**, the output smoothing module **1938** is optional.

As further described below, in the example of FIG. **19**, the down-convert and delay module **1924** and the first and second delay modules **1928** and **1930** include switches that are controlled by a clock having two phases,  $\phi_1$  and  $\phi_2$ .  $\phi_1$  and  $\phi_2$  preferably have the same frequency, and are non-overlapping (alternatively, a plurality such as two clock signals having these characteristics could be used). As used herein, the term “non-overlapping” is defined as two or more signals where only one of the signals is active at any given time. In some embodiments, signals are “active” when they are high. In other embodiments, signals are active when they are low.

Preferably, each of these switches closes on a rising edge of  $\phi_1$  or  $\phi_2$ , and opens on the next corresponding falling edge of  $\phi_1$  or  $\phi_2$ . However, the invention is not limited to this example. As will be apparent to persons skilled in the relevant art(s), other clock conventions can be used to control the switches.

In the example of FIG. **19**, it is assumed that  $\alpha_1$  is equal to one. Thus, the output of the down-convert and delay module **1924** is not scaled. As evident from the embodiments described above, however, the invention is not limited to this example.

The example UDF module **1922** has a filter center frequency of 900.2 MHz and a filter bandwidth of 570 KHz. The pass band of the UDF module **1922** is on the order of 899.915 MHz to 900.485 MHz. The Q factor of the UDF module **1922** is approximately 1879 (i.e., 900.2 MHz divided by 570 KHz).

The operation of the UDF module **1922** shall now be described with reference to a Table **1802** (FIG. **18**) that indicates example values at nodes in the UDF module **1922** at a number of consecutive time increments. It is assumed in Table **1802** that the UDF module **1922** begins operating at time  $t-1$ . As indicated below, the UDF module **1922** reaches steady state a few time units after operation begins. The number of time units necessary for a given UDF module to reach steady state depends on the configuration of the UDF module, and will be apparent to persons skilled in the relevant art(s) based on the teachings contained herein.

At the rising edge of  $\phi_1$  at time  $t-1$ , a switch **1950** in the down-convert and delay module **1924** closes. This allows a capacitor **1952** to charge to the current value of an input signal,  $VI_{t-1}$ , such that node **1902** is at  $VI_{t-1}$ . This is indicated by cell **1804** in FIG. **18**. In effect, the combination of the switch **1950** and the capacitor **1952** in the down-convert and delay module **1924** operates to translate the frequency of the input signal  $VI$  to a desired lower frequency, such as IF or baseband. Thus, the value stored in the capacitor **1952** represents an instance of a down-converted image of the input signal  $VI$ .

The manner in which the down-convert and delay module **1924** performs frequency down-conversion is further described elsewhere in this application, and is additionally described in pending U.S. application “Method and System for Down-Converting Electromagnetic Signals,” Ser. No. 09/176,022, filed Oct. 21, 1998, which is herein incorporated by reference in its entirety.

Also at the rising edge of  $\phi_1$  at time  $t-1$ , a switch **1958** in the first delay module **1928** closes, allowing a capacitor **1960** to charge to  $VO_{t-1}$ , such that node **1906** is at  $VO_{t-1}$ . This is indicated by cell **1806** in Table **1802**. (In practice,  $VO_{t-1}$  is undefined at this point. However, for ease of understanding,  $VO_{t-1}$  shall continue to be used for purposes of explanation.)

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Also at the rising edge of  $\phi_1$  at time  $t-1$ , a switch **1966** in the second delay module **1930** closes, allowing a capacitor **1968** to charge to a value stored in a capacitor **1964**. At this time, however, the value in capacitor **1964** is undefined, so the value in capacitor **1968** is undefined. This is indicated by cell **1807** in table **1802**.

At the rising edge of  $\phi_2$  at time  $t-1$ , a switch **1954** in the down-convert and delay module **1924** closes, allowing a capacitor **1956** to charge to the level of the capacitor **1952**. Accordingly, the capacitor **1956** charges to  $VI_{t-1}$ , such that node **1904** is at  $VI_{t-1}$ . This is indicated by cell **1810** in Table **1802**.

The UDF module **1922** may optionally include a unity gain module **1990A** between capacitors **1952** and **1956**. The unity gain module **1990A** operates as a current source to enable capacitor **1956** to charge without draining the charge from capacitor **1952**. For a similar reason, the UDF module **1922** may include other unity gain modules **1990B-1990G**. It should be understood that, for many embodiments and applications of the invention, these unity gain modules **1990A-1990G** are optional. The structure and operation of the unity gain modules **1990** will be apparent to persons skilled in the relevant art(s).

Also at the rising edge of  $\phi_2$  at time  $t-1$ , a switch **1962** in the first delay module **1928** closes, allowing a capacitor **1964** to charge to the level of the capacitor **1960**. Accordingly, the capacitor **1964** charges to  $VO_{t-1}$ , such that node **1908** is at  $VO_{t-1}$ . This is indicated by cell **1814** in Table **1802**.

Also at the rising edge of  $\phi_2$  at time  $t-1$ , a switch **1970** in the second delay module **1930** closes, allowing a capacitor **1972** to charge to a value stored in a capacitor **1968**. At this time, however, the value in capacitor **1968** is undefined, so the value in capacitor **1972** is undefined. This is indicated by cell **1815** in table **1802**.

At time  $t$ , at the rising edge of  $\phi_1$ , the switch **1950** in the down-convert and delay module **1924** closes. This allows the capacitor **1952** to charge to  $VI_t$ , such that node **1902** is at  $VI_t$ . This is indicated in cell **1816** of Table **1802**.

Also at the rising edge of  $\phi_1$  at time  $t$ , the switch **1958** in the first delay module **1928** closes, thereby allowing the capacitor **1960** to charge to  $VO_t$ . Accordingly, node **1906** is at  $VO_t$ . This is indicated in cell **1820** in Table **1802**.

Further at the rising edge of  $\phi_1$  at time  $t$ , the switch **1966** in the second delay module **1930** closes, allowing a capacitor **1968** to charge to the level of the capacitor **1964**. Therefore, the capacitor **1968** charges to  $VO_{t-1}$ , such that node **1910** is at  $VO_{t-1}$ . This is indicated by cell **1824** in Table **1802**.

At the rising edge of  $\phi_2$  at time  $t$ , the switch **1954** in the down-convert and delay module **1924** closes, allowing the capacitor **1956** to charge to the level of the capacitor **1952**. Accordingly, the capacitor **1956** charges to  $VI_t$ , such that node **1904** is at  $VI_t$ . This is indicated by cell **1828** in Table **1802**.

Also at the rising edge of  $\phi_2$  at time  $t$ , the switch **1962** in the first delay module **1928** closes, allowing the capacitor **1964** to charge to the level in the capacitor **1960**. Therefore, the capacitor **1964** charges to  $VO_t$ , such that node **1908** is at  $VO_t$ . This is indicated by cell **1832** in Table **1802**.

Further at the rising edge of  $\phi_2$  at time  $t$ , the switch **1970** in the second delay module **1930** closes, allowing the capacitor **1972** in the second delay module **1930** to charge to the level of the capacitor **1968** in the second delay module **1930**. Therefore, the capacitor **1972** charges to  $VO_{t-1}$ , such that node **1912** is at  $VO_{t-1}$ . This is indicated in cell **1836** of FIG. **18**.

At time  $t+1$ , at the rising edge of  $\phi_1$ , the switch **1950** in the down-convert and delay module **1924** closes, allowing the

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capacitor **1952** to charge to  $VI_{t+1}$ . Therefore, node **1902** is at  $VI_{t+1}$ , as indicated by cell **1838** of Table **1802**.

Also at the rising edge of  $\phi_1$  at time  $t+1$ , the switch **1958** in the first delay module **1928** closes, allowing the capacitor **1960** to charge to  $VO_{t+1}$ . Accordingly, node **1906** is at  $VO_{t+1}$ , as indicated by cell **1842** in Table **1802**.

Further at the rising edge of  $\phi_1$  at time  $t+1$ , the switch **1966** in the second delay module **1930** closes, allowing the capacitor **1968** to charge to the level of the capacitor **1964**. Accordingly, the capacitor **1968** charges to  $VO_t$ , as indicated by cell **1846** of Table **1802**.

In the example of FIG. **19**, the first scaling module **1932** scales the value at node **1908** (i.e., the output of the first delay module **1928**) by a scaling factor of  $-0.1$ . Accordingly, the value present at node **1914** at time  $t+1$  is  $-0.1*VO_t$ . Similarly, the second scaling module **1934** scales the value present at node **1912** (i.e., the output of the second scaling module **1930**) by a scaling factor of  $-0.8$ . Accordingly, the value present at node **1916** is  $-0.8*VO_{t-1}$  at time  $t+1$ .

At time  $t+1$ , the values at the inputs of the summer **1926** are:  $VI_t$  at node **1904**,  $-0.1*VO_t$  at node **1914**, and  $-0.8*VO_{t-1}$  at node **1916** (in the example of FIG. **19**, the values at nodes **1914** and **1916** are summed by a second summer **1925**, and this sum is presented to the summer **1926**). Accordingly, at time  $t+1$ , the summer generates a signal equal to  $VI_t-0.1*VO_t-0.8*VO_{t-1}$ .

At the rising edge of  $\phi_1$  at time  $t+1$ , a switch **1991** in the output sample and hold module **1936** closes, thereby allowing a capacitor **1992** to charge to  $VO_{t+1}$ . Accordingly, the capacitor **1992** charges to  $VO_{t+1}$ , which is equal to the sum generated by the adder **1926**. As just noted, this value is equal to:  $VI_t-0.1*VO_t-0.8*VO_{t-1}$ . This is indicated in cell **1850** of Table **1802**. This value is presented to the optional output smoothing module **1938**, which smooths the signal to thereby generate the instance of the output signal  $VO_{t+1}$ . It is apparent from inspection that this value of  $VO_{t+1}$  is consistent with the band pass filter transfer function of EQ. **1**.

Further details of unified down-conversion and filtering as described in this section are presented in pending U.S. application "Integrated Frequency Translation And Selectivity," Ser. No. 09/175,966, filed Oct. 21, 1998, incorporated herein by reference in its entirety.

#### 6. Example Application Embodiments of the Invention

As noted above, the UFT module of the present invention is a very powerful and flexible device. Its flexibility is illustrated, in part, by the wide range of applications in which it can be used. Its power is illustrated, in part, by the usefulness and performance of such applications.

Example applications of the UFT module were described above. In particular, frequency down-conversion, frequency up-conversion, enhanced signal reception, and unified down-conversion and filtering applications of the UFT module were summarized above, and are further described below. These applications of the UFT module are discussed herein for illustrative purposes. The invention is not limited to these example applications. Additional applications of the UFT module will be apparent to persons skilled in the relevant art(s), based on the teachings contained herein.

For example, the present invention can be used in applications that involve frequency down-conversion. This is shown in FIG. **1C**, for example, where an example UFT module **115** is used in a down-conversion module **114**. In this capacity, the UFT module **115** frequency down-converts an input signal to an output signal. This is also shown in FIG. **7**, for example, where an example UFT module **706** is part of a down-conversion module **704**, which is part of a receiver **702**.

The present invention can be used in applications that involve frequency up-conversion. This is shown in FIG. 1D, for example, where an example UFT module 117 is used in a frequency up-conversion module 116. In this capacity, the UFT module 117 frequency up-converts an input signal to an output signal. This is also shown in FIG. 8, for example, where an example UFT module 806 is part of up-conversion module 804, which is part of a transmitter 802.

The present invention can be used in environments having one or more transmitters 902 and one or more receivers 906, as illustrated in FIG. 9. In such environments, one or more of the transmitters 902 may be implemented using a UFT module, as shown for example in FIG. 8. Also, one or more of the receivers 906 may be implemented using a UFT module, as shown for example in FIG. 7.

The invention can be used to implement a transceiver. An example transceiver 1002 is illustrated in FIG. 10. The transceiver 1002 includes a transmitter 1004 and a receiver 1008. Either the transmitter 1004 or the receiver 1008 can be implemented using a UFT module. Alternatively, the transmitter 1004 can be implemented using a UFT module 1006, and the receiver 1008 can be implemented using a UFT module 1010. This embodiment is shown in FIG. 10.

Another transceiver embodiment according to the invention is shown in FIG. 11. In this transceiver 1102, the transmitter 1104 and the receiver 1108 are implemented using a single UFT module 1106. In other words, the transmitter 1104 and the receiver 1108 share a UFT module 1106.

As described elsewhere in this application, the invention is directed to methods and systems for enhanced signal reception (ESR). Various ESR embodiments include an ESR module (transmit) in a transmitter 1202, and an ESR module (receive) in a receiver 1210. An example ESR embodiment configured in this manner is illustrated in FIG. 12.

The ESR module (transmit) 1204 includes a frequency up-conversion module 1206. Some embodiments of this frequency up-conversion module 1206 may be implemented using a UFT module, such as that shown in FIG. 1D.

The ESR module (receive) 1212 includes a frequency down-conversion module 1214. Some embodiments of this frequency down-conversion module 1214 may be implemented using a UFT module, such as that shown in FIG. 1C.

As described elsewhere in this application, the invention is directed to methods and systems for unified down-conversion and filtering (UDF). An example unified down-conversion and filtering module 1302 is illustrated in FIG. 13. The unified down-conversion and filtering module 1302 includes a frequency down-conversion module 1304 and a filtering module 1306. According to the invention, the frequency down-conversion module 1304 and the filtering module 1306 are implemented using a UFT module 1308, as indicated in FIG. 13.

Unified down-conversion and filtering according to the invention is useful in applications involving filtering and/or frequency down-conversion. This is depicted, for example, in FIGS. 15A-15F. FIGS. 15A-15C indicate that unified down-conversion and filtering according to the invention is useful in applications where filtering precedes, follows, or both precedes and follows frequency down-conversion. FIG. 15D indicates that a unified down-conversion and filtering module 1524 according to the invention can be utilized as a filter 1522 (i.e., where the extent of frequency down-conversion by the down-converter in the unified down-conversion and filtering module 1524 is minimized). FIG. 15E indicates that a unified down-conversion and filtering module 1528 according to the invention can be utilized as a down-converter 1526 (i.e., where the filter in the unified down-conversion and filtering module 1528 passes substantially all frequencies). FIG. 15F

illustrates that the unified down-conversion and filtering module 1532 can be used as an amplifier. It is noted that one or more UDF modules can be used in applications that involve at least one or more of filtering, frequency translation, and amplification.

For example, receivers, which typically perform filtering, down-conversion, and filtering operations, can be implemented using one or more unified down-conversion and filtering modules. This is illustrated, for example, in FIG. 14.

The methods and systems of unified down-conversion and filtering of the invention have many other applications. For example, as discussed herein, the enhanced signal reception (ESR) module (receive) operates to down-convert a signal containing a plurality of spectrums. The ESR module (receive) also operates to isolate the spectrums in the down-converted signal, where such isolation is implemented via filtering in some embodiments. According to embodiments of the invention, the ESR module (receive) is implemented using one or more unified down-conversion and filtering (UDF) modules. This is illustrated, for example, in FIG. 16. In the example of FIG. 16, one or more of the UDF modules 1610, 1612, 1614 operates to down-convert a received signal. The UDF modules 1610, 1612, 1614 also operate to filter the down-converted signal so as to isolate the spectrum(s) contained therein. As noted above, the UDF modules 1610, 1612, 1614 are implemented using the universal frequency translation (UFT) modules of the invention.

The invention is not limited to the applications of the UFT module described above. For example, and without limitation, subsets of the applications (methods and/or structures) described herein (and others that would be apparent to persons skilled in the relevant art(s) based on the herein teachings) can be associated to forth useful combinations.

For example, transmitters and receivers are two applications of the UFT module. FIG. 10 illustrates a transceiver 1002 that is formed by combining these two applications of the UFT module, i.e., by combining a transmitter 1004 with a receiver 1008.

Also, ESR (enhanced signal reception) and unified down-conversion and filtering are two other applications of the UFT module. FIG. 16 illustrates an example where ESR and unified down-conversion and filtering are combined to forth a modified enhanced signal reception system.

The invention is not limited to the example applications of the UFT module discussed herein. Also, the invention is not limited to the example combinations of applications of the UFT module discussed herein. These examples were provided for illustrative purposes only, and are not limiting. Other applications and combinations of such applications will be apparent to persons skilled in the relevant art(s) based on the teachings contained herein. Such applications and combinations include, for example and without limitation, applications/combinations comprising and/or involving one or more of: (1) frequency translation; (2) frequency down-conversion; (3) frequency up-conversion; (4) receiving; (5) transmitting; (6) filtering; and/or (7) signal transmission and reception in environments containing potentially jamming signals.

Additional example applications are described below.

#### 7. Universal Transmitter

The present invention is directed at a universal transmitter using, in embodiments, two or more UFT modules in a balanced vector modulator configuration. The universal transmitter can be used to create virtually every known and useful waveform used in analog and digital communications applications in wired and wireless markets. By appropriately selecting the inputs to the universal transmitter, a host of

signals can be synthesized including but not limited to AM, FM, BPSK, QPSK, MSK, QAM, OFDM, multi-tone, and spread-spectrum signals (including CDMA and frequency hopping). As will be shown, the universal transmitter can up-convert these waveforms using less components than that seen with conventional super-hetrodyne approaches. In other words, the universal transmitter does not require multiple IF stages (having intermediate filtering) to up-convert complex waveforms that have demanding spectral growth requirements. The elimination of intermediate IF stages reduces part count in the transmitter and therefore leads to cost savings. As will be shown, the present invention achieves these savings without sacrificing performance.

Furthermore, the use of a balanced configuration means that carrier insertion can be attenuated or controlled during up-conversion of a baseband signal. Carrier insertion is caused by the variation of transmitter components (e.g. resistors, capacitors, etc.), which produces DC offset voltages throughout the transmitter. Any DC offset voltage gets up-converted, along with the baseband signal, and generates spectral energy (or carrier insertion) at the carrier frequency  $f_c$ . In many transmit applications, it is highly desirable to minimize the carrier insertion in an up-converted signal because the sideband(s) carry the baseband information and any carrier insertion is wasted energy that reduces efficiency.

FIGS. 25A-B graphically illustrate carrier insertion in the context of up-converted signals that carry baseband information in the corresponding signal sidebands. FIG. 25A depicts an up-converted signal **2502** having minimal carrier energy **2504** when compared to sidebands **2506a** and **2506b**. In these transmitter applications, the present invention can be configured to minimize carrier insertion by limiting the relative DC offset voltage that is present in the transmitter. Alternatively, some transmit applications require sufficient carrier insertion for coherent demodulation of the transmitted signal at the receiver. This illustrated by FIG. 25B, which shows up-converted signal **2508** having carrier energy **2510** that is somewhat larger than sidebands **2512a** and **2512b**. In these applications, the present invention can be configured to introduce a DC offset voltage that generates the desired carrier insertion.

#### 7.1 Universal Transmitter Having 2 UFT Modules

FIG. 26A illustrates a transmitter **2602** according to embodiments of the present invention. Transmitter **2602** includes a balanced modulator/up-converter **2604**, a control signal generator **2642**, an optional filter **2606**, and an optional amplifier **2608**. Transmitter **2602** up-converts a baseband signal **2610** to produce an output signal **2640** that is conditioned for wireless or wire line transmission. In doing so, the balanced modulator **2604** receives the baseband signal **2610** and samples the baseband signal in a differential and balanced fashion to generate a harmonically rich signal **2638**. The harmonically rich signal **2638** includes multiple harmonic images, where each image contains the baseband information in the baseband signal **2610**. The optional bandpass filter **2606** may be included to select a harmonic of interest (or a subset of harmonics) in the signal **2558** for transmission. The optional amplifier **2608** may be included to amplify the selected harmonic prior to transmission. The universal transmitter is further described at a high level by the flowchart **6200** that is shown in FIG. 62. A more detailed structural and operational description of the balanced modulator follows thereafter.

Referring to flowchart **6200**, in step **6202**, the balanced modulator **2604** receives the baseband signal **2610**.

In step **6204**, the balanced modulator **2604** samples the baseband signal in a differential and balanced fashion accord-

ing to a first and second control signals that are phase shifted with respect to each other. The resulting harmonically rich signal **2638** includes multiple harmonic images that repeat at harmonics of the sampling frequency, where each image contains the necessary amplitude and frequency information to reconstruct the baseband signal **2610**.

In embodiments of the invention, the control signals include pulses having pulse widths (or apertures) that are established to improve energy transfer to a desired harmonic of the harmonically rich signal. In further embodiments of the invention, DC offset voltages are minimized between sampling modules as indicated in step **6206**, thereby minimizing carrier insertion in the harmonic images of the harmonically rich signal **2638**.

In step **6208**, the optional bandpass filter **2606** selects the desired harmonic of interest (or a subset of harmonics) in from the harmonically rich signal **2638** for transmission.

In step **6210**, the optional amplifier **2608** amplifies the selected harmonic(s) prior to transmission.

In step **6212**, the selected harmonic(s) is transmitted over a communications medium.

##### 7.1.1 Balanced Modulator Detailed Description

Referring to the example embodiment shown in FIG. 26A, the balanced modulator **2604** includes the following components: a buffer/inverter **2612**; summer amplifiers **2618**, **2619**; UFT modules **2624** and **2628** having controlled switches **2648** and **2650**, respectively; an inductor **2626**; a blocking capacitor **2636**; and a DC terminal **2611**. As stated above, the balanced modulator **2604** differentially samples the baseband signal **2610** to generate a harmonically rich signal **2638**. More specifically, the UFT modules **2624** and **2628** sample the baseband signal in differential fashion according to control signals **2623** and **2627**, respectively. A DC reference voltage **2613** is applied to terminal **2611** and is uniformly distributed to the UFT modules **2624** and **2628**. The distributed DC voltage **2613** prevents any DC offset voltages from developing between the UFT modules, which can lead to carrier insertion in the harmonically rich signal **2638** as described above. The operation of the balanced modulator **2604** is discussed in greater detail with reference to flowchart **6300** (FIG. 63), as follows.

In step **6302**, the buffer/inverter **2612** receives the input baseband signal **2610** and generates input signal **2614** and inverted input signal **2616**. Input signal **2614** is substantially similar to signal **2610**, and inverted signal **2616** is an inverted version of signal **2614**. As such, the buffer/inverter **2612** converts the (single-ended) baseband signal **2610** into differential input signals **2614** and **2616** that will be sampled by the UFT modules. Buffer/inverter **2612** can be implemented using known operational amplifier (op amp) circuits, as will be understood by those skilled in the arts, although the invention is not limited to this example.

In step **6304**, the summer amplifier **2618** sums the DC reference voltage **2613** applied to terminal **2611** with the input signal **2614**, to generate a combined signal **2620**. Likewise, the summer amplifier **2619** sums the DC reference voltage **2613** with the inverted input signal **2616** to generate a combined signal **2622**. Summer amplifiers **2618** and **2619** can be implemented using known op amp summer circuits, and can be designed to have a specified gain or attenuation, including unity gain, although the invention is not limited to this example. The DC reference voltage **2613** is also distributed to the outputs of both UFT modules **2624** and **2628** through the inductor **2626** as is shown.

In step **6306**, the control signal generator **2642** generates control signals **2623** and **2627** that are shown by way of example in FIG. 27B and FIG. 27C, respectively. As illus-

trated, both control signals **2623** and **2627** have the same period  $T_S$  as a master clock signal **2645** (FIG. 27A), but have a pulse width (or aperture) of  $T_A$ . In the example, control signal **2623** triggers on the rising pulse edge of the master clock signal **2645**, and control signal **2627** triggers on the falling pulse edge of the master clock signal **2645**. Therefore, control signals **2623** and **2627** are shifted in time by 180 degrees relative to each other. In embodiments of invention, the master clock signal **2645** (and therefore the control signals **2623** and **2627**) have a frequency that is a sub-harmonic of the desired output signal **2640**. The invention is not limited to the example of FIGS. 27A-27C.

In one embodiment, the control signal generator **2642** includes an oscillator **2646**, pulse generators **2644a** and **2644b**, and an inverter **2647** as shown. In operation, the oscillator **2646** generates the master clock signal **2645**, which is illustrated in FIG. 27A as a periodic square wave having pulses with a period of  $T_S$ . Other clock signals could be used including but not limited to sinusoidal waves, as will be understood by those skilled in the arts. Pulse generator **2644a** receives the master clock signal **2645** and triggers on the rising pulse edge, to generate the control signal **2623**. Inverter **2647** inverts the clock signal **2645** to generate an inverted clock signal **2643**. The pulse generator **2644b** receives the inverted clock signal **2643** and triggers on the rising pulse edge (which is the falling edge of clock signal **2645**), to generate the control signal **2627**.

FIG. 74A-E illustrate example embodiments for the pulse generator **2644**. FIG. 74A illustrates a pulse generator **7402**. The pulse generator **7402** generates pulses **7408** having pulse width  $T_A$  from an input signal **7404**. Example input signals **7404** and pulses **7408** are depicted in FIGS. 74B and 74C, respectively. The input signal **7404** can be any type of periodic signal, including, but not limited to, a sinusoid, a square wave, a saw-tooth wave etc. The pulse width (or aperture)  $T_A$  of the pulses **7408** is determined by delay **7406** of the pulse generator **7402**. The pulse generator **7402** also includes an optional inverter **7410**, which is optionally added for polarity considerations as understood by those skilled in the arts. The example logic and implementation shown for the pulse generator **7402** is provided for illustrative purposes only, and is not limiting. The actual logic employed can take many forms. Additional examples of pulse generation logic are shown in FIGS. 74D and 74E. FIG. 74D illustrates a rising edge pulse generator **7412** that triggers on the rising edge of input signal **7404**. FIG. 74E illustrates a falling edge pulse generator **7416** that triggers on the falling edge of the input signal **7404**.

In step **6308**, the UFT module **2624** samples the combined signal **2620** according to the control signal **2623** to generate harmonically rich signal **2630**. More specifically, the switch **2648** closes during the pulse widths  $T_A$  of the control signal **2623** to sample the combined signal **2620** resulting in the harmonically rich signal **2630**. FIG. 26B illustrates an exemplary frequency spectrum for the harmonically rich signal **2630** having harmonic images **2652a-n**. The images **2652** repeat at harmonics of the sampling frequency  $1/T_S$ , at infinitum, where each image **2652** contains the necessary amplitude, frequency, and phase information to reconstruct the baseband signal **2610**. As discussed further below, the relative amplitude of the frequency images is generally a function of the harmonic number and the pulse width  $T_A$ . As such, the relative amplitude of a particular harmonic **2652** can be increased (or decreased) by adjusting the pulse width  $T_A$  of the control signal **2623**. In general, shorter pulse widths of  $T_A$  shift more energy into the higher frequency harmonics, and longer pulse widths of  $T_A$  shift energy into the lower frequency harmonics. The generation of harmonically rich sig-

nals by sampling an input signal according to a controlled aperture have been described earlier in this application in the section titled, "Frequency Up-conversion Using Universal Frequency Translation", and is illustrated by FIGS. 3-6. A more detailed discussion of frequency up-conversion using a switch with a controlled sampling aperture is discussed in the co-pending patent application titled, "Method and System for Frequency Up-Conversion," Ser. No. 09/176,154, filed on Oct. 21, 1998, and incorporated herein by reference.

In step **6310**, the UFT module **2628** samples the combined signal **2622** according to the control signal **2627** to generate harmonically rich signal **2634**. More specifically, the switch **2650** closes during the pulse widths  $T_A$  of the control signal **2627** to sample the combined signal **2622** resulting in the harmonically rich signal **2634**. The harmonically rich signal **2634** includes multiple frequency images of baseband signal **2610** that repeat at harmonics of the sampling frequency ( $1/T_S$ ), similar to that for the harmonically rich signal **2630**. However, the images in the signal **2634** are phase-shifted compared to those in signal **2630** because of the inversion of signal **2616** compared to signal **2614**, and because of the relative phase shift between the control signals **2623** and **2627**.

In step **6312**, the node **2632** sums the harmonically rich signals **2632** and **2634** to generate harmonically rich signal **2633**. FIG. 26C illustrates an exemplary frequency spectrum for the harmonically rich signal **2633** that has multiple images **2654a-n** that repeat at harmonics of the sampling frequency  $1/T_S$ . Each image **2654** includes the necessary amplitude, frequency and phase information to reconstruct the baseband signal **2610**. The capacitor **2636** operates as a DC blocking capacitor and substantially passes the harmonics in the harmonically rich signal **2633** to generate harmonically rich signal **2638** at the output of the modulator **2604**.

In step **6208**, the optional filter **2606** can be used to select a desired harmonic image for transmission. This is represented for example by a passband **2656** that selects the harmonic image **2654c** for transmission in FIG. 26C.

An advantage of the modulator **2604** is that it is fully balanced, which substantially minimizes (or eliminates) any DC voltage offset between the two UFT modules **2624** and **2628**. DC offset is minimized because the reference voltage **2613** contributes a consistent DC component to the input signals **2620** and **2622** through the summing amplifiers **2618** and **2619**, respectively. Furthermore, the reference voltage **2613** is also directly coupled to the outputs of the UFT modules **2624** and **2628** through the inductor **2626** and the node **2632**. The result of controlling the DC offset between the UFT modules is that carrier insertion is minimized in the harmonic images of the harmonically rich signal **2638**. As discussed above, carrier insertion is substantially wasted energy because the information for a modulated signal is carried in the sidebands of the modulated signal and not in the carrier. Therefore, it is often desirable to minimize the energy at the carrier frequency by controlling the relative DC offset.

#### 7.1.2 Balanced Modulator Example Signal Diagrams and Mathematical Description

In order to further describe the invention, FIGS. 27D-27I illustrate various example signal diagrams (vs. time) that are representative of the invention. These signal diagrams are meant for example purposes only and are not meant to be limiting. FIG. 27D illustrates a signal **2702** that is representative of the input baseband signal **2610** (FIG. 26A). FIG. 27E illustrates a step function **2704** that is an expanded portion of the signal **2702** from time  $t_0$  to  $t_1$ , and represents signal **2614** at the output of the buffer/inverter **2612**. Similarly, FIG. 27F illustrates a signal **2706** that is an inverted version of the

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signal 2704, and represents the signal 2616 at the inverted output of buffer/inverter 2612. For analysis purposes, a step function is a good approximation for a portion of a single bit of data (for the baseband signal 2610) because the clock rates of the control signals 2623 and 2627 are significantly higher than the data rates of the baseband signal 2610. For example, if the data rate is in the KHz frequency range, then the clock rate will preferably be in MHz frequency range in order to generate an output signal in the GHz frequency range.

Still referring to FIGS. 27D-I, FIG. 27G illustrates a signal 2708 that an example of the harmonically rich signal 2630 when the step function 2704 is sampled according to the control signal 2623 in FIG. 27B. The signal 2708 includes positive pulses 2709 as referenced to the DC voltage 2613. Likewise, FIG. 27H illustrates a signal 2710 that is an example of the harmonically rich signal 2634 when the step function 2706 is sampled according to the control signal 2627. The signal 2710 includes negative pulses 2711 as referenced to the DC voltage 2613, which are time-shifted relative to the positive pulses 2709 in signal 2708.

Still referring to FIGS. 27D-I, the FIG. 27I illustrates a signal 2712 that is the combination of signal 2708 (FIG. 27G) and the signal 2710 (FIG. 27H), and is an example of the harmonically rich signal 2633 at the output of the summing node 2632. As illustrated, the signal 2712 spends approximately as much time above the DC reference voltage 2613 as below the DC reference voltage 2613 over a limited time period. For example, over a time period 2714, the energy in the positive pulses 2709a-b is canceled out by the energy in the negative pulses 2711a-b. This is indicative of minimal (or zero) DC offset between the UFT modules 2624 and 2628, which results in minimal carrier insertion during the sampling process.

Still referring to FIG. 27I, the time axis of the signal 2712 can be phased in such a manner to represent the waveform as an odd function. For such an arrangement, the Fourier series is readily calculated to obtain:

$$I_C(t) = \sum_{n=1}^{\infty} \left( \frac{4 \sin\left(\frac{n\pi T_A}{T_S}\right) \cdot \left(\frac{n\pi}{2}\right)}{n\pi} \right) \cdot \sin\left(\frac{2n\pi t}{T_S}\right). \quad \text{Equation 1}$$

where:

$T_S$ =period of the master clock 2645

$T_A$ =pulse width of the control signals 2623 and 2627

$n$ =harmonic number

As shown by Equation 1, the relative amplitude of the frequency images is generally a function of the harmonic number  $n$ , and the ratio of  $T_A/T_S$ . As indicated, the  $T_A/T_S$  ratio represents the ratio of the pulse width of the control signals relative to the period of the sub-harmonic master clock. The  $T_A/T_S$  ratio can be optimized in order to maximize the amplitude of the frequency image at a given harmonic. For example, if a passband waveform is desired to be created at 5x the frequency of the sub-harmonic clock, then a baseline power for that harmonic extraction may be calculated for the fifth harmonic ( $n=5$ ) as:

$$I_C(t) = \left( \frac{4 \sin\left(\frac{5\pi T_A}{T_S}\right)}{5\pi} \right) \cdot \sin(5\omega_S t). \quad \text{Equation 2}$$

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As shown by Equation 2,  $I_C(t)$  for the fifth harmonic is a sinusoidal function having an amplitude that is proportional to the  $\sin(5\pi T_A/T_S)$ . The signal amplitude can be maximized by setting  $T_A=(1/10 \cdot T_S)$  so that  $\sin(5\pi T_A/T_S)=\sin(\pi/2)=1$ . Doing so results in the equation:

$$I_C(t)|_{n=5} = \frac{4}{5\pi} (\sin(5\omega_S t)). \quad \text{Equation 3}$$

This component is a frequency at 5x of the sampling frequency of sub-harmonic clock, and can be extracted from the Fourier series via a bandpass filter (such as bandpass filter 2606) that is centered around  $5f_S$ . The extracted frequency component can then be optionally amplified by the amplifier 2608 prior to transmission on a wireless or wire-line communications channel or channels.

Equation 3 can be extended to reflect the inclusion of a message signal as illustrated by equation 4 below:

$$m(t) \cdot I_C(t)|_{\substack{n=5 \\ \theta=\theta(t)}} = \frac{4 \cdot m(t)}{5\pi} (\sin(5\omega_S t + 5\theta(t))). \quad \text{Equation 4}$$

Equation 4 illustrates that a message signal can be carried in harmonically rich signals 2633 such that both amplitude and phase can be modulated. In other words,  $m(t)$  is modulated for amplitude and  $\theta(t)$  is modulated for phase. In such cases, it should be noted that  $\theta(t)$  is augmented modulo  $n$  while the amplitude modulation  $m(t)$  is simply scaled. Therefore, complex waveforms may be reconstructed from their Fourier series with multiple aperture UFT combinations.

As discussed above, the signal amplitude for the 5th harmonic was maximized by setting the sampling aperture width  $T_A=1/10 T_S$ , where  $T_S$  is the period of the master clock signal. This can be restated and generalized as setting  $T_A=1/2$  the period (or  $\pi$  radians) at the harmonic of interest. In other words, the signal amplitude of any harmonic  $n$  can be maximized by sampling the input waveform with a sampling aperture of  $T_A=1/2$  the period of the harmonic of interest ( $n$ ). Based on this discussion, it is apparent that varying the aperture changes the harmonic and amplitude content of the output waveform. For example, if the sub-harmonic clock has a frequency of 200 MHz, then the fifth harmonic is at 1 GHz. The amplitude of the fifth harmonic is maximized by setting the aperture width  $T_A=500$  picoseconds, which equates to  $1/2$  the period (or  $\pi$  radians) at 1 GHz.

FIG. 27J depicts a frequency plot 2716 that graphically illustrates the effect of varying the sampling aperture of the control signals on the harmonically rich signal 2633 given a 200 MHz harmonic clock. The frequency plot 2716 compares two frequency spectrums 2718 and 2720 for different control signal apertures given a 200 MHz clock. More specifically, the frequency spectrum 2718 is an example spectrum for signal 2633 given the 200 MHz clock with the aperture  $T_A=500$  psec (where 500 psec is  $\pi$  radians at the 5th harmonic of 1 GHz). Similarly, the frequency spectrum 2720 is an example spectrum for signal 2633 given a 200 MHz clock that is a square wave (so  $T_A=5000$  psec). The spectrum 2718 includes multiple harmonics 2718a-i, and the frequency spectrum 2720 includes multiple harmonics 2720a-e. [It is noted that spectrum 2720 includes only the odd harmonics as predicted by Fourier analysis for a square wave.] At 1 GHz (which is the 5th harmonic), the signal amplitude of the two frequency spectrums 2718e and 2720c are approximately

equal. However, at 200 MHz, the frequency spectrum **2718a** has a much lower amplitude than the frequency spectrum **2720a**, and therefore the frequency spectrum **2718** is more efficient than the frequency spectrum **2720**, assuming the desired harmonic is the 5th harmonic. In other words, assuming 1 GHz is the desired harmonic, the frequency spectrum **2718** wastes less energy at the 200 MHz fundamental than does the frequency spectrum **2718**.

#### 7.1.3 Balanced Modulator Having a Shunt Configuration

FIG. **56A** illustrates a universal transmitter **5600** that is a second embodiment of a universal transmitter having two balanced UFT modules in a shunt configuration. (In contrast, the balanced modulator **2604** can be described as having a series configuration based on the orientation of the UFT modules.) Transmitter **5600** includes a balanced modulator **5601**, the control signal generator **2642**, the optional bandpass filter **2606**, and the optional amplifier **2608**. The transmitter **5600** up-converts a baseband signal **5602** to produce an output signal **5636** that is conditioned for wireless or wire line transmission. In doing so, the balanced modulator **5601** receives the baseband signal **5602** and shunts the baseband signal to ground in a differential and balanced fashion to generate a harmonically rich signal **5634**. The harmonically rich signal **5634** includes multiple harmonic images, where each image contains the baseband information in the baseband signal **5602**. In other words, each harmonic image includes the necessary amplitude, frequency, and phase information to reconstruct the baseband signal **5602**. The optional bandpass filter **2606** may be included to select a harmonic of interest (or a subset of harmonics) in the signal **5634** for transmission. The optional amplifier **2608** may be included to amplify the selected harmonic prior to transmission, resulting in the output signal **5636**.

The balanced modulator **5601** includes the following components: a buffer/inverter **5604**; optional impedances **5610**, **5612**; UFT modules **5616** and **5622** having controlled switches **5618** and **5624**, respectively; blocking capacitors **5628** and **5630**; and a terminal **5620** that is tied to ground. As stated above, the balanced modulator **5601** differentially shunts the baseband signal **5602** to ground, resulting in a harmonically rich signal **5634**. More specifically, the UFT modules **5616** and **5622** alternately shunt the baseband signal to terminal **5620** according to control signals **2623** and **2627**, respectively. Terminal **5620** is tied to ground and prevents any DC offset voltages from developing between the UFT modules **5616** and **5622**. As described above, a DC offset voltage can lead to undesired carrier insertion. The operation of the balanced modulator **5601** is described in greater detail according to the flowchart **6400** (FIG. **64**) as follows.

In step **6402**, the buffer/inverter **5604** receives the input baseband signal **5602** and generates I signal **5606** and inverted I signal **5608**. I signal **5606** is substantially similar to the baseband signal **5602**, and the inverted I signal **5608** is an inverted version of signal **5602**. As such, the buffer/inverter **5604** converts the (single-ended) baseband signal **5602** into differential signals **5606** and **5608** that are sampled by the UFT modules. Buffer/inverter **5604** can be implemented using known operational amplifier (op amp) circuits, as will be understood by those skilled in the arts, although the invention is not limited to this example.

In step **6404**, the control signal generator **2642** generates control signals **2623** and **2627** from the master clock signal **2645**. Examples of the master clock signal **2645**, control signal **2623**, and control signal **2627** are shown in FIGS. **27A-C**, respectively. As illustrated, both control signals **2623** and **2627** have the same period  $T_s$  as a master clock signal

**2645**, but have a pulse width (or aperture) of  $T_A$ . Control signal **2623** triggers on the rising pulse edge of the master clock signal **2645**, and control signal **2627** triggers on the falling pulse edge of the master clock signal **2645**. Therefore, control signals **2623** and **2627** are shifted in time by 180 degrees relative to each other. A specific embodiment of the control signal generator **2642** is illustrated in FIG. **26A**, and was discussed in detail above.

In step **6406**, the UFT module **5616** shunts the signal **5606** to ground according to the control signal **2623**, to generate a harmonically rich signal **5614**. More specifically, the switch **5618** closes and shorts the signal **5606** to ground (at terminal **5620**) during the aperture width  $T_A$  of the control signal **2623**, to generate the harmonically rich signal **5614**. FIG. **56B** illustrates an exemplary frequency spectrum for the harmonically rich signal **5618** having harmonic images **5650a-n**. The images **5650** repeat at harmonics of the sampling frequency  $1/T_s$ , at infinity, where each image **5650** contains the necessary amplitude, frequency, and phase information to reconstruct the baseband signal **5602**. The generation of harmonically rich signals by sampling an input signal according to a controlled aperture have been described earlier in this application in the section titled, "Frequency Up-conversion Using Universal Frequency Translation", and is illustrated by FIGS. **3-6**. A more detailed discussion of frequency up-conversion using a switch with a controlled sampling aperture is discussed in the co-pending patent application titled, "Method and System for Frequency Up-Conversion," Ser. No. 09/176,154, filed on Oct. 21, 1998, and incorporated herein by reference.

The relative amplitude of the frequency images **5650** is generally a function of the harmonic number and the pulse width  $T_A$ . As such, the relative amplitude of a particular harmonic **5650** can be increased (or decreased) by adjusting the pulse width  $T_A$  of the control signal **2623**. In general, shorter pulse widths of  $T_A$  shift more energy into the higher frequency harmonics, and longer pulse widths of  $T_A$  shift energy into the lower frequency harmonics. Additionally, the relative amplitude of a particular harmonic **5650** can also be adjusted by adding/tuning an optional impedance **5610**. Impedance **5610** operates as a filter that emphasizes a particular harmonic in the harmonically rich signal **5614**.

In step **6408**, the UFT module **5622** shunts the inverted signal **5608** to ground according to the control signal **2627**, to generate a harmonically rich signal **5626**. More specifically, the switch **5624** closes during the pulse widths  $T_A$  and shorts the inverted I signal **5608** to ground (at terminal **5620**), to generate the harmonically rich signal **5626**. At any given time, only one of input signals **5606** or **5608** is shorted to ground because the pulses in the control signals **2623** and **2627** are phase shifted with respect to each other, as shown in FIGS. **27B** and **27C**.

The harmonically rich signal **5626** includes multiple frequency images of baseband signal **5602** that repeat at harmonics of the sampling frequency ( $1/T_s$ ), similar to that for the harmonically rich signal **5614**. However, the images in the signal **5626** are phase-shifted compared to those in signal **5614** because of the inversion of the signal **5608** compared to the signal **5606**, and because of the relative phase shift between the control signals **2623** and **2627**. The optional impedance **5612** can be included to emphasize a particular harmonic of interest, and is similar to the impedance **5610** above.

In step **6410**, the node **5632** sums the harmonically rich signals **5614** and **5626** to generate the harmonically rich signal **5634**. The capacitors **5628** and **5630** operate as blocking capacitors that substantially pass the respective harmoni-

cally rich signals **5614** and **5626** to the node **5632**. (The capacitor values may be chosen to substantially block baseband frequency components as well.) FIG. **56C** illustrates an exemplary frequency spectrum for the harmonically rich signal **5634** that has multiple images **5652<sub>a-n</sub>** that repeat at harmonics of the sampling frequency  $1/T_S$ . Each image **5652** includes the necessary amplitude, frequency, and phase information to reconstruct the baseband signal **5602**. The optional filter **2606** can be used to select the harmonic image of interest for transmission. This is represented by a passband **5656** that selects the harmonic image **5632<sub>c</sub>** for transmission.

An advantage of the modulator **5601** is that it is fully balanced, which substantially minimizes (or eliminates) any DC voltage offset between the two UFT modules **5612** and **5614**. DC offset is minimized because the UFT modules **5616** and **5622** are both connected to ground at terminal **5620**. The result of controlling the DC offset between the UFT modules is that carrier insertion is minimized in the harmonic images of the harmonically rich signal **5634**. As discussed above, carrier insertion is substantially wasted energy because the information for a modulated signal is carried in the sidebands of the modulated signal and not in the carrier. Therefore, it is often desirable to minimize the energy at the carrier frequency by controlling the relative DC offset.

#### 7.1.4 Balanced Modulator FET Configuration

As described above, the balanced modulators **2604** and **5601** utilize two balanced UFT modules to sample the input baseband signals to generate harmonically rich signals that contain the up-converted baseband information. More specifically, the UFT modules include controlled switches that sample the baseband signal in a balanced and differential fashion. FIGS. **26D** and **56D** illustrate embodiments of the controlled switch in the UFT module.

FIG. **26D** illustrates an example embodiment of the modulator **2604** (FIG. **26B**) where the controlled switches in the UFT modules are field effect transistors (FET). More specifically, the controlled switches **2648** and **2628** are embodied as FET **2658** and FET **2660**, respectively. The FET **2658** and **2660** are oriented so that their gates are controlled by the control signals **2623** and **2627**, so that the control signals control the FET conductance. For the FET **2658**, the combined baseband signal **2620** is received at the source of the FET **2658** and is sampled according to the control signal **2623** to produce the harmonically rich signal **2630** at the drain of the FET **2658**. Likewise, the combined baseband signal **2622** is received at the source of the FET **2660** and is sampled according to the control signal **2627** to produce the harmonically rich signal **2634** at the drain of FET **2660**. The source and drain orientation that is illustrated is not limiting, as the source and drains can be switched for most FETs. In other words, the combined baseband signal can be received at the drain of the FETs, and the harmonically rich signals can be taken from the source of the FETs, as will be understood by those skilled in the relevant arts.

FIG. **56D** illustrates an embodiment of the modulator **5600** (FIG. **56**) where the controlled switches in the UFT modules are field effect transistors (FET). More specifically, the controlled switches **5618** and **5624** are embodied as FET **5636** and FET **5638**, respectively. The FETs **5636** and **5638** are oriented so that their gates are controlled by the control signals **2623** and **2627**, respectively, so that the control signals determine FET conductance. For the FET **5636**, the baseband signal **5606** is received at the source of the FET **5636** and shunted to ground according to the control signal **2623**, to produce the harmonically rich signal **5614**. Likewise, the baseband signal **5608** is received at the source of the FET **5638** and is shunted to grounding according to the control

signal **2627**, to produce the harmonically rich signal **5626**. The source and drain orientation that is illustrated is not limiting, as the source and drains can be switched for most FETs, as will be understood by those skilled in the relevant arts.

#### 7.1.5 Universal Transmitter Configured for Carrier Insertion

As discussed above, the transmitters **2602** and **5600** have a balanced configuration that substantially eliminates any DC offset and results in minimal carrier insertion in the output signal **2640**. Minimal carrier insertion is generally desired for most applications because the carrier signal carries no information and reduces the overall transmitter efficiency. However, some applications require the received signal to have sufficient carrier energy for the receiver to extract the carrier for coherent demodulation. In support thereof, the present invention can be configured to provide the necessary carrier insertion by implementing a DC offset between the two sampling UFT modules.

FIG. **28A** illustrates a transmitter **2802** that up-converts a baseband signal **2806** to an output signal **2822** having carrier insertion. As is shown, the transmitter **2802** is similar to the transmitter **2602** (FIG. **26A**) with the exception that the up-converter/modulator **2804** is configured to accept two DC reference voltages. In contrast, modulator **2604** was configured to accept only one DC reference voltage. More specifically, the modulator **2804** includes a terminal **2809** to accept a DC reference voltage **2808**, and a terminal **2813** to accept a DC reference voltage **2814**. Vr **2808** appears at the UFT module **2624** though summer amplifier **2618** and the inductor **2810**. Vr **2814** appears at UFT module **2628** through the summer amplifier **2619** and the inductor **2816**. Capacitors **2812** and **2818** operate as blocking capacitors. If Vr **2808** is different from Vr **2814** then a DC offset voltage will be exist between UFT module **2624** and UFT module **2628**, which will be up-converted at the carrier frequency in the harmonically rich signal **2820**. More specifically, each harmonic image in the harmonically rich signal **2820** will include a carrier signal as depicted in FIG. **28B**.

FIG. **28B** illustrates an exemplary frequency spectrum for the harmonically rich signal **2820** that has multiple harmonic images **2824<sub>a-n</sub>**. In addition to carrying the baseband information in the sidebands, each harmonic image **2824** also includes a carrier signal **2826** that exists at respective harmonic of the sampling frequency  $1/T_S$ . The amplitude of the carrier signal increases with increasing DC offset voltage. Therefore, as the difference between Vr **2808** and Vr **2814** widens, the amplitude of each carrier signal **2826** increases. Likewise, as the difference between Vr **2808** and Vr **2814** shrinks, the amplitude of each carrier signal **2826** shrinks. As with transmitter **2802**, the optional bandpass filter **2606** can be included to select a desired harmonic image for transmission. This is represented by passband **2828** in FIG. **28B**.

#### 7.2 Universal Transmitter in I Q Configuration:

As described above, the balanced modulators **2604** and **5601** up-convert a baseband signal to a harmonically rich signal having multiple harmonic images of the baseband information. By combining two balanced modulators, IQ configurations can be formed for up-converting I and Q baseband signals. In doing so, either the (series type) balanced modulator **2604** or the (shunt type) balanced modulator can be utilized. IQ modulators having both series and shunt configurations are described below.

##### 7.2.1 IQ Transmitter Using Series-Type Balanced Modulator

FIG. **29** illustrates an IQ transmitter **2920** with an in-phase (I) and quadrature (Q) configuration according to embodi-

ments of the invention. The transmitter 2920 includes an IQ balanced modulator 2910, an optional filter 2914, and an optional amplifier 2916. The transmitter 2920 is useful for transmitting complex I Q waveforms and does so in a balanced manner to control DC offset and carrier insertion. In doing so, the modulator 2910 receives an I baseband signal 2902 and a Q baseband signal 2904 and up-converts these signals to generate a combined harmonically rich signal 2912. The harmonically rich signal 2912 includes multiple harmonics images, where each image contains the baseband information in the I signal 2902 and the Q signal 2904. The optional bandpass filter 2914 may be included to select a harmonic of interest (or subset of harmonics) from the signal 2912 for transmission. The optional amplifier 2916 may be included to amplify the selected harmonic prior to transmission, to generate the IQ output signal 2918.

As stated above, the balanced IQ modulator 2910 up-converts the I baseband signal 2902 and the Q baseband signal 2904 in a balanced manner to generate the combined harmonically rich signal 2912 that carries the I and Q baseband information. To do so, the modulator 2910 utilizes two balanced modulators 2604 from FIG. 26A, a signal combiner 2908, and a DC terminal 2907. The operation of the balanced modulator 2910 and other circuits in the transmitter is described according to the flowchart 6500 in FIG. 65, as follows.

In step 6502, the IQ modulator 2910 receives the I baseband signal 2902 and the Q baseband signal 2904.

In step 6504, the I balanced modulator 2604a samples the I baseband signal 2902 in a differential fashion using the control signals 2623 and 2627 to generate a harmonically rich signal 2911a. The harmonically rich signal 2911a contains multiple harmonic images of the I baseband information, similar to the harmonically rich signal 2630 in FIG. 26B.

In step 6506, the balanced modulator 2604b samples the Q baseband signal 2904 in a differential fashion using control signals 2623 and 2627 to generate harmonically rich signal 2911b, where the harmonically rich signal 2911b contains multiple harmonic images of the Q baseband signal 2904. The operation of the balanced modulator 2604 and the generation of harmonically rich signals was fully described above and illustrated in FIGS. 26A-C, to which the reader is referred for further details.

In step 6508, the DC terminal 2907 receives a DC voltage 2906 that is distributed to both modulators 2604a and 2604b. The DC voltage 2906 is distributed to both the input and output of both UFT modules 2624 and 2628 in each modulator 2604. This minimizes (or prevents) DC offset voltages from developing between the four UFT modules, and thereby minimizes or prevents any carrier insertion during the sampling steps 6504 and 6506.

In step 6510, the 90 degree signal combiner 2908 combines the harmonically rich signals 2911a and 2911b to generate IQ harmonically rich signal 2912. This is further illustrated in FIGS. 30A-C. FIG. 30A depicts an exemplary frequency spectrum for the harmonically rich signal 2911a having harmonic images 3002a-n. The images 3002 repeat at harmonics of the sampling frequency  $1/T_s$ , where each image 3002 contains the necessary amplitude and frequency information to reconstruct the I baseband signal 2902. Likewise, FIG. 30B depicts an exemplary frequency spectrum for the harmonically rich signal 2911b having harmonic images 3004a-n. The harmonic images 3004a-n also repeat at harmonics of the sampling frequency  $1/T_s$ , where each image 3004 contains the necessary amplitude, frequency, and phase information to reconstruct the Q baseband signal 2904. FIG. 30C illustrates an exemplary frequency spectrum for the combined harmoni-

cally rich signal 2912 having images 3006. Each image 3006 carries the I baseband information and the Q baseband information from the corresponding images 3002 and 3004, respectively, without substantially increasing the frequency bandwidth occupied by each harmonic 3006. This can occur because the signal combiner 2908 phase shifts the Q signal 2911b by 90 degrees relative to the I signal 2911a. The result is that the images 3002a-n and 3004a-n effectively share the signal bandwidth do to their orthogonal relationship. For example, the images 3002a and 3004a effectively share the frequency spectrum that is represented by the image 3006a.

In step 6512, the optional filter 2914 can be included to select a harmonic of interest, as represented by the passband 3008 selecting the image 3006c in FIG. 30c.

In step 6514, the optional amplifier 2916 can be included to amplify the harmonic (or harmonics) of interest prior to transmission.

In step 6516, the selected harmonic (or harmonics) is transmitted over a communications medium.

FIG. 31A illustrates a transmitter 3108 that is a second embodiment for an I Q transmitter having a balanced configuration. Transmitter 3108 is similar to the transmitter 2920 except that the 90 degree phase shift between the I and Q channels is achieved by phase shifting the control signals instead of using a 90 degree signal combiner to combine the harmonically rich signals. More specifically, delays 3104a and 3104b delay the control signals 2623 and 2627 for the Q channel modulator 2604b by 90 degrees relative to the control signals for the I channel modulator 2604a. As a result, the Q modulator 2604b samples the Q baseband signal 2904 with 90 degree delay relative to the sampling of the I baseband signal 2902 by the I channel modulator 2604a. Therefore, the Q harmonically rich signal 2911b is phase shifted by 90 degrees relative to the I harmonically rich signal. Since the phase shift is achieved using the control signals, an in-phase signal combiner 3106 combines the harmonically rich signals 2911a and 2911b, to generate the harmonically rich signal 2912.

FIG. 31B illustrates a transmitter 3118 that is similar to transmitter 3108 in FIG. 31A. The difference being that the transmitter 3118 has a modulator 3120 that utilizes a summing node 3122 to sum the signals 2911a and 2911b instead of the in-phase signal combiner 3106 that is used in modulator 3102 of transmitter 3108.

FIG. 55A-55D illustrate various detailed circuit implementations of the transmitter 2920 in FIG. 29. These circuit implementations are meant for example purposes only, and are not meant to be limiting.

FIG. 55A illustrates I input circuitry 5502a and Q input circuitry 5502b that receive the I and Q input signals 2902 and 2904, respectively.

FIG. 55B illustrates the I channel circuitry 5506 that processes an I data 5504a from the I input circuit 5502a.

FIG. 55C illustrates the Q channel circuitry 5508 that processes the Q data 5504b from the Q input circuit 5502b.

FIG. 55D illustrates the output combiner circuit 5512 that combines the I channel data 5507 and the Q channel data 5510 to generate the output signal 2918.

#### 7.2.2. IQ Transmitter Using Shunt-Type Balanced Modulator

FIG. 57 illustrates an IQ transmitter 5700 that is another IQ transmitter embodiment according to the present invention. The transmitter 5700 includes an IQ balanced modulator 5701, an optional filter 5712, and an optional amplifier 5714. During operation, the modulator 5701 up-converts an I baseband signal 5702 and a Q baseband signal 5704 to generate a combined harmonically rich signal 5711. The harmonically

rich signal **5711** includes multiple harmonics images, where each image contains the baseband information in the I signal **5702** and the Q signal **5704**. The optional bandpass filter **5712** may be included to select a harmonic of interest (or subset of harmonics) from the harmonically rich signal **5711** for transmission. The optional amplifier **5714** may be included to amplify the selected harmonic prior to transmission, to generate the IQ output signal **5716**.

The IQ modulator **5701** includes two balanced modulators **5601** from FIG. **56**, and a 90 degree signal combiner **5710** as shown. The operation of the IQ modulator **5701** is described in reference to the flowchart **6600** (FIG. **66**), as follows. The order of the steps in flowchart **6600** is not limiting.

In step **6602**, the balanced modulator **5701** receives the I baseband signal **5702** and the Q baseband signal **5704**.

In step **6604**, the balanced modulator **5601a** differentially shunts the I baseband signal **5702** to ground according to the control signals **2623** and **2627**, to generate a harmonically rich signal **5706**. More specifically, the UFT modules **5616a** and **5622a** alternately shunt the I baseband signal and an inverted version of the I baseband signal to ground according to the control signals **2623** and **2627**, respectively. The operation of the balanced modulator **5601** and the generation of harmonically rich signals was fully described above and is illustrated in FIGS. **56A-C**, to which the reader is referred for further details. As such, the harmonically rich signal **5706** contains multiple harmonic images of the I baseband information as described above.

In step **6606**, the balanced modulator **5601b** differentially shunts the Q baseband signal **5704** to ground according to control signals **2623** and **2627**, to generate harmonically rich signal **5708**. More specifically, the UFT modules **5616b** and **5622b** alternately shunt the Q baseband signal and an inverted version of the Q baseband signal to ground, according to the control signals **2623** and **2627**, respectively. As such, the harmonically rich signal **5708** contains multiple harmonic images that contain the Q baseband information.

In step **6608**, the 90 degree signal combiner **5710** combines the harmonically rich signals **5706** and **5708** to generate IQ harmonically rich signal **5711**. This is further illustrated in FIGS. **58A-C**. FIG. **58A** depicts an exemplary frequency spectrum for the harmonically rich signal **5706** having harmonic images **5802a-n**. The harmonic images **5802** repeat at harmonics of the sampling frequency  $1/T_s$ , where each image **5802** contains the necessary amplitude, frequency, and phase information to reconstruct the I baseband signal **5702**. Likewise, FIG. **58B** depicts an exemplary frequency spectrum for the harmonically rich signal **5708** having harmonic images **5804a-n**. The harmonic images **5804a-n** also repeat at harmonics of the sampling frequency  $1/T_s$ , where each image **5804** contains the necessary amplitude, frequency, and phase information to reconstruct the Q baseband signal **5704**.

FIG. **58C** illustrates an exemplary frequency spectrum for the IQ harmonically rich signal **5711** having images **5806a-n**. Each image **5806** carries the I baseband information and the Q baseband information from the corresponding images **5802** and **5804**, respectively, without substantially increasing the frequency bandwidth occupied by each image **5806**. This can occur because the signal combiner **5710** phase shifts the Q signal **5708** by 90 degrees relative to the I signal **5706**.

In step **6610**, the optional filter **5712** may be included to select a harmonic of interest, as represented by the passband **5808** selecting the image **5806c** in FIG. **58C**.

In step **6612**, the optional amplifier **5714** can be included to amplify the selected harmonic image **5806** prior to transmission.

In step **6614**, the selected harmonic (or harmonics) is transmitted over a communications medium.

FIG. **59** illustrates a transmitter **5900** that is another embodiment for an I Q transmitter having a balanced configuration. Transmitter **5900** is similar to the transmitter **5700** except that the 90 degree phase shift between the I and Q channels is achieved by phase shifting the control signals instead of using a 90 degree signal combiner to combine the harmonically rich signals. More specifically, delays **5904a** and **5904b** delay the control signals **2623** and **2627** for the Q channel modulator **5601b** by 90 degrees relative the control signals for the I channel modulator **5601a**. As a result, the Q modulator **5601b** samples the Q baseband signal **5704** with a 90 degree delay relative to the sampling of the I baseband signal **5702** by the I channel modulator **5601a**. Therefore, the Q harmonically rich signal **5708** is phase shifted by 90 degrees relative to the I harmonically rich signal **5706**. Since the phase shift is achieved using the control signals, an in-phase signal combiner **5906** combines the harmonically rich signals **5706** and **5708**, to generate the harmonically rich signal **5711**.

FIG. **60** illustrates a transmitter **6000** that is similar to transmitter **5900** in FIG. **59**. The difference being that the transmitter **6000** has a balanced modulator **6002** that utilizes a summing node **6004** to sum the I harmonically rich signal **5706** and the Q harmonically rich signal **5708** instead of the in-phase signal combiner **5906** that is used in the modulator **5902** of transmitter **5900**. The 90 degree phase shift between the I and Q channels is implemented by delaying the Q clock signals using 90 degree delays **5904**, as shown.

#### 7.2.3 IQ Transmitters Configured for Carrier Insertion

The transmitters **2920** (FIG. **29**) and **3108** (FIG. **31A**) have a balanced configuration that substantially eliminates any DC offset and results in minimal carrier insertion in the IQ output signal **2918**. Minimal carrier insertion is generally desired for most applications because the carrier signal carries no information and reduces the overall transmitter efficiency. However, some applications require the received signal to have sufficient carrier energy for the receiver to extract the carrier for coherent demodulation. In support thereof, FIG. **32** illustrates a transmitter **3202** to provide any necessary carrier insertion by implementing a DC offset between the two sets of sampling UFT modules.

Transmitter **3202** is similar to the transmitter **2920** with the exception that a modulator **3204** in transmitter **3202** is configured to accept two DC reference voltages so that the I channel modulator **2604a** can be biased separately from the Q channel modulator **2604b**. More specifically, modulator **3204** includes a terminal **3206** to accept a DC voltage reference **3207**, and a terminal **3208** to accept a DC voltage reference **3209**. Voltage **3207** biases the UFT modules **2624a** and **2628a** in the I channel modulator **2604a**. Likewise, voltage **3209** biases the UFT modules **2624b** and **2628b** in the Q channel modulator **2604b**. When voltage **3207** is different from voltage **3209**, then a DC offset will appear between the I channel modulator **2604a** and the Q channel modulator **2604b**, which results in carrier insertion in the IQ harmonically rich signal **2912**. The relative amplitude of the carrier frequency energy increases in proportion to the amount of DC offset.

FIG. **33** illustrates a transmitter **3302** that is a second embodiment of an IQ transmitter having two DC terminals to cause DC offset, and therefore carrier insertion. Transmitter **3302** is similar to transmitter **3202** except that the 90 degree phase shift between the I and Q channels is achieved by phase shifting the control signals, similar to that done in transmitter **3108**. More specifically, delays **3304a** and **3304b** phase shift the control signals **2623** and **2627** for the Q channel modula-

tor **2604b** relative to those of the I channel modulator **2604a**. As a result, the Q modulator **2604b** samples the Q baseband signal **2904** with 90 degree delay relative to the sampling of the I baseband signal **2902** by the I channel modulator **2604a**. Therefore, the Q harmonically rich signal **2911b** is phase shifted by 90 degrees relative to the I harmonically rich signal, which is then combined by the in-phase combiner **3306**.

### 7.3 Universal Transmitter and CDMA

The universal transmitter **2920** (FIG. **29**) and the universal transmitter **5700** (FIG. **57**) can be used to up-convert every known useful analog and digital baseband waveform including but not limited to: AM, FM, PM, BPSK, QPSK, MSK, QAM, OFDM, multi-tone, and spread spectrum signals. For further illustration, FIG. **34A** and FIG. **34B** depict transmitter **2920** configured to up-convert the mentioned modulation waveforms. FIG. **34A** illustrates transmitter **2920** configured to up-convert non-complex waveform including AM and shaped BPSK. In FIG. **34A**, these non-complex (and non-IQ) waveforms are received on the I terminal **3402**, and the Q input **3404** is grounded since only a single channel is needed. FIG. **34B** illustrates a transmitter **2920** that is configured to receive both I and Q inputs for the up-conversion of complex waveforms including QPSK, QAM, OFDM, GSM, and spread spectrum waveforms (including CDMA and frequency hopping). The transmitters in FIGS. **34A** and **34B** are presented for illustrative purposes, and are not limiting. Other embodiments are possible, as will be appreciated in view of the teachings herein.

CDMA is an input waveform that is of particular interest for communications applications. CDMA is the fastest growing digital cellular communications standard in many regions, and now is widely accepted as the foundation for the competing third generation (3G) wireless standard. CDMA is considered to be the among the most demanding of the current digital cellular standards in terms of RF performance requirements.

#### 7.3.1 IS-95 CDMA Specifications

FIG. **35A** and FIG. **35B** illustrate the CDMA specifications for base station and mobile transmitters as required by the IS-95 standard. FIG. **35A** illustrates a base station CDMA signal **3502** having a main lobe **3504** and sidelobes **3506a** and **3506b**. For base station transmissions, IS-95 requires that the sidelobes **3506a,b** are at least 45 dB below the mainlobe **3504** (or 45 dbc) at an offset frequency of 750 kHz, and 60 dbc at an offset frequency of 1.98 MHz. FIG. **35B** illustrates similar requirements for a mobile CDMA signal **3508** having a main lobe **3510** and sidelobes **3512a** and **3512b**. For mobile transmissions, CDMA requires that the sidelobes **3512a,b** are at least 42 dbc at a frequency offset of 885 kHz, and 54 dbc at a frequency offset 1.98 MHz.

Rho is another well known performance parameter for CDMA. Rho is a figure-of-merit that measures the amplitude and phase distortion of a CDMA signal that has been processed in some manner (e.g. amplified, up-converted, filtered, etc.) The maximum theoretical value for Rho is 1.0, which indicates no distortion during the processing of the CDMA signal. The IS-95 requirement for the baseband-to-RF interface is Rho=0.9912. As will be shown by the test results below, the transmitter **2920** (in FIG. **29**) can up-convert a CDMA baseband signal and achieve Rho values of approximately Rho=0.9967. Furthermore, the modulator **2910** in the transmitter **2920** achieves these results in standard CMOS (although the invention is not limited to this example implementation), without doing multiple up-conversions and IF filtering that is associated with conventional super-heterodyne configurations.

#### 7.3.2 Conventional CDMA Transmitter

Before describing the CDMA implementation of transmitter **2920**, it is useful to describe a conventional super-heterodyne approach that is used to meet the IS-95 specifications. FIG. **36** illustrates a conventional CDMA transmitter **3600** that up-converts an input signal **3602** to an output CDMA signal **3634**. The conventional CDMA transmitter **3600** includes: a baseband processor **3604**, a baseband filter **3608**, a first mixer **3612**, an amplifier **3616**, a SAW filter **3620**, a second mixer **3624**, a power amplifier **3628**, and a band-select filter **3632**. The conventional CDMA transmitter operates as follows.

The baseband processor **3604** spreads the input signal **3602** with I and Q spreading codes to generate I signal **3606a** and Q signal **3606b**, which are consistent with CDMA IS-95 standards. The baseband filter **3608** filters the signals **3606** with the aim of reducing the sidelobes so as to meet the sidelobe specifications that were discussed in FIGS. **35A** and **35B**. Mixer **3612** up-converts the signal **3610** using a first LO signal **3613** to generate an IF signal **3614**. IF amplifier **3616** amplifies the IF signal **3614** to generate IF signal **3618**. SAW filter **3620** has a bandpass response that filters the IF signal **3618** to suppress any sidelobes caused by the non-linear operations of the mixer **3614**. As is understood by those skilled in the arts, SAW filters provide significant signal suppression outside the passband, but are relatively expensive and large compared to other transmitter components. Furthermore, SAW filters are typically built on specialized materials that cannot be integrated onto a standard CMOS chip with other components. Mixer **3624** up-converts the signal **3622** using a second LO signal **3625** to generate RF signal **3626**. Power amplifier **3628** amplifies RF signal **3626** to generate signal **3630**. Band-select filter **3632** bandpass filters RF signal **3630** to suppress any unwanted harmonics in output signal **3634**.

It is noted that transmitter **3602** up-converts the input signal **3602** using an IF chain **3636** that includes the first mixer **3612**, the amplifier **3616**, the SAW filter **3620**, and the second mixer **3624**. The IF chain **3636** up-converts the input signal to an IF frequency and does IF amplification and SAW filtering in order to meet the IS-95 sidelobe and figure-of-merit specifications. This is done because conventional wisdom teaches that a CDMA baseband signal cannot be up-converted directly from baseband to RF, and still meet the IS-95 linearity requirements.

#### 7.3.3 CDMA Transmitter Using the Present Invention

For comparison, FIG. **37A** illustrates an example CDMA transmitter **3700** according to embodiments of the present invention. The CDMA transmitter **3700** includes (it is noted that the invention is not limited to this example): the baseband processor **3604**; the baseband filter **3608**; the IQ modulator **2910** (from FIG. **29**), the control signal generator **2642**, the sub-harmonic oscillator **2646**, the power amplifier **3628**, and the filter **3632**. In the example of FIG. **37A**, the baseband processor **3604**, baseband filter **3608**, amplifier **3628**, and the band-select filter **3632** are the same as that used in the conventional transmitter **3602** in FIG. **36**. The difference is that the IQ modulator **2910** in transmitter **3700** completely replaces the IF chain **3636** in the conventional transmitter **3602**. This is possible because the modulator **2910** up-converts a CDMA signal directly from baseband-to-RF without any IF processing. The detailed operation of the CDMA transmitter **3700** is described with reference to the flowchart **7300** (FIG. **73**) as follows.

In step **7302**, the input baseband signal **3702** is received.

In step **7304**, the CDMA baseband processor **3604** receives the input signal **3702** and spreads the input signal **3702** using

I and Q spreading codes, to generate an I signal **3704a** and a Q signal **3704b**. As will be understood, the I spreading code and Q spreading codes can be different to improve isolation between the I and Q channels.

In step **7306**, the baseband filter **3608** bandpass filters the I signal **3704a** and the Q signal **3704b** to generate filtered I signal **3706a** and filtered Q signal **3706b**. As mentioned above, baseband filtering is done to improve sidelobe suppression in the CDMA output signal.

FIGS. **37B-37D** illustrate the effect of the baseband filter **3608** on the I and Q inputs signals. FIG. **37B** depicts multiple signal traces (over time) for the filtered I signal **3706a**, and FIG. **37C** depicts multiple signal traces for the filtered Q signal **3706b**. As shown, the signals **3706a,b** can be described as having an “eyelid” shape having a thickness **3715**. The thickness **3715** reflects the steepness of passband roll off of the baseband filter **3608**. In other words, a relatively thick eyelid in the time domain reflects a steep passband roll off in the frequency domain, and results in lower sidelobes for the output CDMA signal. However, there is a tradeoff, because as the eyelids become thicker, then there is a higher probability that channel noise will cause a logic error during decoding at the receiver. The voltage rails **3714** represent the +1/-1 logic states for the I and Q signals **3706**, and correspond to the logic states in complex signal space that are shown in FIG. **37D**.

In step **7308**, the IQ modulator **2910** samples I and Q input signals **3706A**, **3706B** in a differential and balanced fashion according to sub-harmonic clock signals **2623** and **2627**, to generate a harmonically rich signal **3708**. FIG. **37E** illustrates the harmonically rich signal **3708** that includes multiple harmonic images **3716a-n** that repeat at harmonics of the sampling frequency  $1/T_s$ . Each image **3716a-n** is a spread spectrum signal that contains the necessary amplitude, frequency, and phase information to reconstruct the input baseband signal **3702**.

In step **7310**, the amplifier **3628** amplifies the harmonically rich signal **3708** to generate an amplified harmonically rich signal **3710**.

Finally, the band-select filter **3632** selects the harmonic of interest from signal **3710**, to generate an CDMA output signal **3712** that meets IS-95 CDMA specifications. This is represented by passband **3718** selecting harmonic image **3716b** in FIG. **37E**.

An advantage of the CDMA transmitter **3700** is in that the modulator **2910** up-converts a CDMA input signal directly from baseband to RF without any IF processing, and still meets the IS-95 sidelobe and figure-of-merit specifications. In other words, the modulator **2910** is sufficiently linear and efficient during the up-conversion process that no IF filtering or amplification is required to meet the IS-95 requirements. Therefore, the entire IF chain **3636** can be replaced by the modulator **2910**, including the expensive SAW filter **3620**. Since the SAW filter is eliminated, substantial portions of the transmitter **3702** can be integrated onto a single CMOS chip, for example, that uses standard CMOS process. More specifically, and for illustrative purposes only, the baseband processor **3604**, the baseband filter **3608**, the modulator **2910**, the oscillator **2646**, and the control signal generator **2642** can be integrated on a single CMOS chip, as illustrated by CMOS chip **3802** in FIG. **38**, although the invention is not limited to this implementation example.

FIG. **37F** illustrates a transmitter **3720** that is similar to transmitter **3700** (FIG. **37A**) except that modulator **5701** replaces the modulator **2910**. Transmitter **3700** operates similar to the transmitter **3700** and has all the same advantages of the transmitter **3700**.

Other embodiments discussed or suggested herein can be used to implement other CDMA transmitters according to the invention.

#### 7.3.4 CDMA Transmitter Measured Test Results

As discussed above, the UFT-based modulator **2910** directly up-converts baseband CDMA signals to RF without any IF filtering, while maintaining the required figures-of-merit for IS-95. The modulator **2910** has been extensively tested in order to specifically determine the performance parameters when up-converting CDMA signals. The test system and measurement results are discussed as follows.

FIG. **39** illustrates a test system **3900** that measures the performance of the modulator **2910** when up-converting CDMA baseband signals. The test system **3900** includes: a Hewlett Packerd (HP) generator E4433B, attenuators **3902a** and **3902b**, control signal generator **2642**, UFT-based modulator **2910**, amplifier/filter module **3904**, cable/attenuator **3906**, and HP 4406A test set. The HP generator E4433B generates I and Q CDMA baseband waveforms that meet the IS-95 test specifications. The waveforms are routed to the UFT-based modulator **2910** through the 8-dB attenuators **3902a** and **3902b**. The HP generator E4433B also generates the sub-harmonic clock signal **2645** that triggers the control signal generator **2642**, where the sub-harmonic clock **2645** has a frequency of 279 MHz. The modulator **2910** up-converts the I and Q baseband signals to generate a harmonic rich signal **3903** having multiple harmonic images that represent the input baseband signal and repeat at the sampling frequency. The amplifier/filter module **3904** selects and amplifies the 3rd harmonic (of the 279 MHz clock signal) in the signal **3903** to generate the signal **3905** at 837 MHz. The HP 4406A test set accepts the signal **3905** for analysis through the cable/attenuator **3906**. The HP 4406A measures CDMA modulation attributes including: Rho, EVM, phase error, amplitude error, output power, carrier insertion, and ACPR. In addition, the signal is demodulated and Walsh code correlation parameters are analyzed. Both forward and reverse links have been characterized using pilot, access, and traffic channels. For further illustration, FIGS. **40-60Z** display the measurement results for the RF spectrum **3905** based on various base station and mobile waveforms that are generated by the HP E443B generator.

FIGS. **40** and **41** summarize the performance parameters of the modulator **2910** as measured by the test set **3900** for base station and mobile station input waveforms, respectively. For the base station, table **4002** includes lists performance parameters that were measured at a base station middle frequency and includes: Rho, EVM, phase error, magnitude error, carrier insertion, and output power. It is noted that Rho=0.997 for the base station middle frequency and exceeds the IS-95 requirement of Rho=0.912. For the mobile station, FIG. **41** illustrates a table **4102** that lists performance parameters that were measured at low, middle, and high frequencies. It is noted that the Rho exceeds the IS-95 requirement (0.912) for each of the low, middle, high frequencies of the measured waveform.

FIG. **42** illustrates a base station constellation **4202** measured during a pilot channel test. A signal constellation plots the various logic combinations for the I and Q signals in complex signal space, and is the raw data for determining the performance parameters (including Rho) that are listed in Table **40**. The performance parameters (in table **40**) are also indicated beside the constellation measurement **4202** for convenience. Again, it is noted that Rho=0.997 for this test. A value of 1 is perfect, and 0.912 is required by the IS-95 CDMA specification, although most manufactures strive for

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values greater than 0.94. This is a remarkable result since the modulator 2910 up-converts directly from baseband-to-RF without any IF filtering.

FIG. 43 illustrates a base station sampled constellation 4302, and depicts the tight constellation samples that are associated with FIG. 42. The symmetry and sample scatter compactness are illustrative of the superior performance of the modulator 2910.

FIG. 44 illustrates a mobile station constellation 4402 measured during an access channel test. As shown,  $\text{Rho}=0.997$  for the mobile station waveforms. Therefore, the modulator 2910 operates very well with conventional and offset shaped QPSK modulation schemes.

FIG. 45 illustrates a mobile station sampled constellation 4502. Constellation 4502 illustrates excellent symmetry for the constellation sample scatter diagram.

FIG. 46 illustrates a base station constellation 4602 using only the HP test equipment. The modulator 2910 has been removed so that the base station signal travels only through the cables that connect the HP signal generator E4433B to the HP 4406A test set. Therefore, constellation 4602 measures signal distortion caused by the test set components (including the cables and the attenuators). It is noted that  $\text{Rho}=0.9994$  for this measurement using base station waveforms. Therefore, at least part of the minimal signal distortion that is indicated in FIGS. 42 and 43 is caused by the test set components, as would be expected by those skilled in the relevant arts.

FIG. 47 illustrates a mobile station constellation 4702 using only the HP test equipment. As in FIG. 46, the modulator 2910 has been removed so that the mobile station signal travels only through the cables that connect the HP signal generator E4433B to the HP 4406A test set. Therefore, constellation 4602 measures signal distortion caused by the test set components (including the cables and the attenuators). It is noted that  $\text{Rho}=0.9991$  for this measurement using mobile station waveforms. Therefore, at least part of the signal distortion indicated in FIGS. 44 and 45 is caused by the test set components, as would be expected.

FIG. 48 illustrates a frequency spectrum 4802 of the signal 3905 with a base station input waveform. The frequency spectrum 4802 has a main lobe and two sidelobes, as expected for a CDMA spread spectrum signal. The adjacent channel power ratio (ACPR) measures the spectral energy at a particular frequency of the side lobes relative to the main lobe. As shown, the frequency spectrum 4802 has an ACPR=-48.34 dBc and -62.18 dBc at offset frequencies of 750 KHz and 1.98 MHz, respectively. The IS-95 ACPR requirement for a base station waveform is -45 dBc and -60 dBc maximum, at the offset frequencies of 750 kHz and 1.98 MHz, respectively. Therefore, the modulator 2910 has more than 3 dB and 2 dB of margin over the IS-95 requirements for the 750 kHz and 1.98 MHz offsets, respectively.

FIG. 49 illustrates a histogram 4902 that corresponds to the spectrum plot in FIG. 48. The histogram 4902 illustrates the distribution of the spectral energy in the signal 3905 for a base station waveform.

FIG. 50 illustrates a frequency spectrum 5002 of the signal 3905 with a mobile station input waveform. As shown, the ACPR measurement is -52.62 dBc and -60.96 dBc for frequency offsets of 885 kHz and 1.98 MHz, respectively. The IS-95 ACPR requirement for a mobile station waveform is approximately -42 dBc and -54 dBc, respectively. Therefore, the modulator 2910 has over 10 dB and 6 dB of margin above the IS-95 requirements for the 885 kHz and 1.98 MHz frequency offsets, respectively.

FIG. 51 illustrates a histogram 5102 that corresponds to the mobile station spectrum plot in FIG. 50. The histogram 5102

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illustrates the distribution of the spectral energy in the signal 3905 for a mobile station waveform.

FIG. 52A illustrates a histogram 5202 for crosstalk vs. CDMA channel with a base station input waveform. More specifically, the HP E4406A was utilized as a receiver to analyze the orthogonality of codes superimposed on the base station modulated spectrum. The HP E4406A demodulated the signal provided by the modulator/transmitter and determined the crosstalk to non-active CDMA channels. The pilot channel is in slot '0' and is the active code for this test. All non-active codes are suppressed in the demodulation process by greater than 40 dB. The IS-95 requirement is 27 dB of suppression so that there is over 13 dB of margin. This implies that the modulator 2910 has excellent phase and amplitude linearity.

In addition to the measurements described above, measurements were also conducted to obtain the timing and phase delays associated with a base station transmit signal composed of pilot and active channels. Delta measurements were extracted with the pilot signal as a reference. The delay and phase are -5.7 ns (absolute) and 7.5 milli radians, worst case. The standard requires less than 50 ns (absolute) and 50 milli radians, which the modulator 2910 exceeded with a large margin.

The performance sensitivity of modulator 2910 was also measured over multiple parameter variations. More specifically, the performance sensitivity was measured vs. IQ input signal level variation and LO signal level variation, for both base station and mobile station modulation schemes. (LO signal level is the signal level of the sub-harmonic clock 2645 in FIG. 39.) FIGS. 52B-O depict performance sensitivity of the modulator 2910 using the base station modulation scheme, and FIGS. 52P-Z depict performance sensitivity using the mobile station modulation scheme. These plots reveal that the modulator 2910 is expected to enable good production yields since there is a large acceptable operating performance range for I/Q and LO peak to peak voltage inputs. The plots are described further as follows.

FIG. 52B illustrates Rho vs. shaped IQ input signal level using base station modulation.

FIG. 52C illustrates transmitted channel power vs. shaped IQ input signal level using base station modulation.

FIG. 52D illustrates ACPR vs. shaped IQ Input signal level using base station modulation.

FIG. 52E illustrates EVM and Magnitude error vs shaped IQ input level using base station modulation.

FIG. 52F illustrates carrier feed thru vs. shaped IQ input signal level using base station modulation.

FIG. 52G illustrates Rho vs. LO signal level using base station modulation.

FIG. 52H illustrates transmitted channel power vs. LO signal level using base station modulation.

FIG. 52I illustrates ACPR vs. LO signal level using base station modulation.

FIG. 52J illustrates EVM and magnitude error vs LO signal level using base station modulation.

FIG. 52K illustrates carrier feed thru vs. LO signal level using base station modulation.

FIG. 52L illustrates carrier feed thru vs IQ input level over a wide range using base station modulation.

FIG. 52M illustrates ACPR vs. shaped IQ input signal level using base station modulation.

FIG. 52N illustrates Rho vs. shaped IQ input signal level using base station modulation.

FIG. 52O illustrates EVM, magnitude error, and phase error vs. shaped IQ input signal level using base station modulation.

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FIG. 52P illustrates Rho vs. shaped IQ input signal level using mobile station modulation.

FIG. 52Q illustrates transmitted channel power vs. shaped IQ input signal level using mobile station modulation.

FIG. 52R illustrates ACPR vs. shaped IQ Input signal level using mobile station modulation.

FIG. 52S illustrates EVM, magnitude error, and phase error vs. shaped IQ input level using mobile station modulation.

FIG. 52T illustrates carrier feed thru vs. shaped I Q input signal level using mobile station modulation.

FIG. 52U illustrates Rho vs. LO signal level using mobile station modulation.

FIG. 52V illustrates transmitted channel power vs. LO signal level using mobile station modulation.

FIG. 52W illustrates ACPR vs. LO signal level using mobile station modulation.

FIG. 52X illustrates EVM and magnitude error vs. LO signal level using mobile station modulation.

FIG. 52Y illustrates carrier feed thru vs. LO signal level using mobile station modulation.

FIG. 52Z illustrates an approximate power budget for a CDMA modulator based on the modulator 2910.

FIGS. 52B-Z illustrate that the UFT-based complex modulator 2910 comfortably exceeds the IS-95 transmitter performance requirements for both mobile and base station modulations, even with signal level variations. Testing indicates that Rho as well as carrier feed through and ACPR are not overly sensitive to variations in I/Q levels and LO levels. Estimated power consumption for the modulator 2910 is lower than equivalent two-state superheterodyne architecture. This means that a practical UFT based CDMA transmitter can be implemented in bulk CMOS and efficiently produced in volume.

The UFT architecture achieves the highest linearity per milliwatt of power consumed of any radio technology of which the inventors are aware. This efficiency comes without a performance penalty, and due to the inherent linearity of the UFT technology, several important performance parameters may actually be improved when compared to traditional transmitter techniques.

Since the UFT technology can be implemented in standard CMOS, new system partitioning options are available that have not existed before. As an example, since the entire UFT-based modulator can be implemented in CMOS, it is plausible that the modulator and other transmitter functions can be integrated with the digital baseband processor leaving only a few external components such as the final bandpass filter and the power amplifier. In addition to the UFT delivering the required linearity and dynamic range performance, the technology also has a high level of immunity to digital noise that would be found on the same substrate when integrated with other digital circuitry. This is a significant step towards enabling a complete wireless system-on-chip solution.

It is noted that the test setup, procedures, and results discussed above and shown in the figures were provided for illustrative purposes only, and do not limit the invention to any particular embodiment, implementation or application.

8.0 Integrated Up-Conversion And Spreading Of A Baseband Signal

Previous sections focused on up-converting a spread spectrum signal directly from baseband-to-RF, without preforming any IF processing. In these embodiments, the baseband signal was already a spread spectrum signal prior to up-conversion. The following discussion focuses on embodiments that perform the spreading function and the frequency translation function in a simultaneously and in an integrated

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manner. One type of spreading code is Code Division Multiple Access (or CDMA), although the invention is not limited to this. The present invention can be implemented in CDMA, and other spread spectrum systems as will be understood by those skilled in the arts based on the teachings herein.

8.1 Integrated Up-Conversion and Spreading Using an Amplitude Shaper

FIG. 53A illustrates a spread spectrum transmitter 5300 that is based on the UFT-based modulator 2604 that was discussed in FIG. 26A. Spread spectrum transmitter 5300 performs simultaneous up-conversion and spreading of an input baseband signal 5302 to generate an output signal 5324. As will shown, the spreading is accomplished by placing the spreading code on the control signals that operate the UFT modules in the modulator 2604 so that the spreading and up-conversion are accomplished in an integrated manner. In order to limit sidelobe spectral growth in the output signal 5324, the amplitude of the input baseband signal 5302 is shaped so as to correspond with the spreading code. The operation of spread spectrum transmitter 5300 is described in detail as follows with reference to flowchart 6700 that is shown in FIG. 67. The order of the steps in flowchart 6700 are not limiting and may be re-arranged as will be understood by those skilled in the arts. (This is generally true of all flowcharts discussed herein).

In step 6701, the spread spectrum transmitter 5300 receives the input baseband signal 5302.

In step 6702, the oscillator 2646 generates the clock signal 2645. As described earlier, the clock signal 2645 is in embodiment a sub-harmonic of the output signal 5324. Furthermore, in embodiments of the invention, the clock signal 2645 is a periodic square wave or sinusoidal clock signal.

In step 6704, a spreading code generator 5314 generates a spreading code 5316. In embodiments of the invention, the spreading code 5316 is a PN code, or any other type of spreading code that is useful for generating spread spectrum signals.

In step 6706, the multiplier 5318 modulates the clock signal 2645 with the spreading code 5316 to generate spread clock signal 5320. As such, the spread clock signal 5320 carries the spreading code 5316.

In step 6708, the control signal generator 2642 receives the spread clock signal 5320, and generates control signals 5321 and 5322 that operate the UFT modules in the modulator 2604. The control signals 5321 and 5322 are similar to clock signals 2623 and 2627 that were discussed in FIG. 26. In other words, the clock signals 5321 and 5322 include a plurality of pulses having a pulse width  $T_A$  that is established to improve energy transfer to a desired harmonic in the resulting harmonically rich signal. Additionally, the control signals 5321 and 5322 are phase shifted with respect to each other by approximately 180 degrees (although the invention is not limited to this example), as were the control signals 2623 and 2627. However, the control signals 5321 and 5322 are modulated with (and carry) the spreading code 5316 because they were generated from spread clock signal 5320.

In step 6710, the amplitude shaper 5304 receives the input baseband signal 5302 and shapes the amplitude so that it corresponds with the spreading code 5316 that is generated by the code generator 5314, resulting in a shaped input signal 5306. This is achieved by feeding the spreading code 5316 back to the amplitude shaper 5304 and smoothing the amplitude of the input baseband signal 5302, accordingly.

FIG. 53B illustrates the resulting shaped input signal 5306 and the corresponding spreading code 5316. The amplitude of the input signal 5302 is shaped such that it is smooth and so that it has zero crossings that are in time synchronization with

the spreading code **5316**. By smoothing input signal amplitude, high frequency components are removed from the input signal prior to sampling, which results lower sidelobe energy in the harmonic images produced during sampling. Implementation of amplitude shaper **5304** will be apparent to persons skilled in the art base on the functional teachings combined herein.

In step **6712**, the low pass filter **5308** filters the shaped input signal **5306** to remove any unwanted high frequency components, resulting in a filtered signal **5310**.

In step **6714**, the modulator **2604** samples the signal **5310** in a balanced and differential manner according to the control signals **5320** and **5322**, to generate a harmonically rich signal **5312**. As discussed in reference to FIG. 26, the control signals **5320** and **5322** trigger the controlled switches in the modulator **2604**, resulting in multiple harmonic images of the baseband signal **5302** in the harmonically rich signal **5312**. Since the control signals carry the spreading code **5316**, the modulator **2604** up-converts and spreads the filtered signal **5310** in an integrated manner during the sampling process. As such, the harmonic images in the harmonically rich signal **5312** are spread spectrum signals. FIG. 53C illustrates the harmonically rich signal **5312** that includes multiple harmonic images **5320a-n** that repeat at harmonics of the sampling frequency  $1/T_s$ . Each image **5320a-n** is a spread spectrum signal that contains the necessary amplitude and frequency information to reconstruct the input baseband signal **5302**.

In step **6716**, the optional filter **2606** selects a desired harmonic (or harmonics) from the harmonically rich signal **5312**. This is presented by the passband **5322** selecting the spread harmonic **5320c** in FIG. 53C.

In step **6718**, the optional amplifier **2608** amplifies the desired harmonic (or harmonics) for transmission.

As mentioned above, an advantage of the spread spectrum transmitter **5300** is that the spreading and up-conversion is accomplished in a simultaneous and integrated manner. This is a result of modulating the control signals that operate the UFT modules in the balanced modulator **2604** with the spreading code prior to sampling of the baseband signal. Furthermore, by shaping the amplitude of the baseband signal prior to sampling, the sidelobe energy in the spread spectrum harmonics is minimized. As discussed above, minimal sidelobe energy is desirable in order to meet the sidelobe standards of the CDMA IS-95 standard (see FIGS. 43A and 43B).

FIG. 61 illustrates an IQ spread spectrum modulator **6100** that is based on the spread spectrum transmitter **5300**. Spread spectrum modulator **6100** performs simultaneous up-conversion and spreading of an I baseband signal **6102** and a Q baseband signal **6118** to generate an output signal **6116** that carries both the I and Q baseband information. The operation of the modulator **6100** is described in detail with reference to the flowchart **6800** that is shown in FIGS. 68A and 68B. The steps in flowchart **6800** are not limiting and may be re-arranged as will be understood by those skilled in the arts.

In step **6801**, the IQ modulator **6100** receives the I data signal **6102** and the Q data signal **6118**.

In step **6802**, the oscillator **2646** generates the clock signal **2645**. As described earlier, the clock signal **2645** is in embodiments a sub-harmonic of the output signal **6116**. Furthermore, in embodiments of the invention, the clock signal **2645** is a periodic square wave or sinusoidal clock signal.

In step **6804**, an I spreading code generator **6140** generates an I spreading code **6144** for the I channel. Likewise, a Q spreading code generator **6138** generates a Q spreading code **6142** for the Q channel. In embodiments of the invention, the spreading codes are PN codes, or any other type of spreading code that is useful for generating spread spectrum signals. In

embodiments of the invention, the I spreading code and Q spreading code can be the same spreading code. Alternatively, the I and Q spreading codes can be different to improve isolation between the I and Q channels, as will be understood by those skilled in the arts.

In step **6806**, the multiplier **5318a** modulates the clock signal **2645** with the I spreading code **6144** to generate a spread clock signal **6136**. Likewise, the multiplier **5318b** modulates the clock signal **2645** with the Q spreading code **6142** to generate a spread clock signal **6134**.

In step **6808**, the control signal generator **2642a** receives the I clock signal **6136** and generates control signals **6130** and **6132** that operate the UFT modules in the modulator **2604a**. The control signals **6130** and **6132** are similar to clock signals **2623** and **2627** that were discussed in FIG. 26. The difference being that signals **6130** and **6132** are modulated with (and carry) the I spreading code **6144**. Likewise, the control signal generator **2642b** receives the Q clock signal **6134** and generates control signals **6126** and **6128** that operate the UFT modules in the modulator **2604b**. In step **6810**, the amplitude shaper **5304a** receives the I data signal **6102** and the shapes the amplitude so that it corresponds with the spreading code **6144**, resulting in I shaped data signal **6104**. This is achieved by feeding the spreading code **6144** back to the amplitude shaper **5304a**. The amplitude shaper then shapes the amplitude of the input baseband signal **6102** to correspond to the spreading code **6144**, as described for spread spectrum transmitter **5300**. More specifically, the amplitude of the input signal **6102** is shaped such that it is smooth and so that it has zero crossings that are in time synchronization with the I spreading code **6144**. Likewise, the amplitude shaper **5304b** receives the Q data signal **6118** and shapes the amplitude of the Q data signal **6118** so that it corresponds with the Q spreading code **6142**, resulting in Q shaped data signal **6120**.

In step **6812**, the low pass filter **5308a** filters the I shaped data signal **6104** to remove any unwanted high frequency components, resulting in a I filtered signal **6106**. Likewise, the low pass filter **5308b** filters the Q shaped data signal **6120**, resulting in Q filtered signal **6122**.

In step **6814**, the modulator **2604a** samples the I filtered signal **6106** in a balanced and differential manner according to the control signals **6130** and **6132**, to generate a harmonically rich signal **6108**. As discussed in reference to FIG. 26, the control signals **6130** and **6132** trigger the controlled switches in the modulator **2604a**, resulting in multiple harmonic images in the harmonically rich signal **6108**, where each image contains the I baseband information. Since the control signals **6130** and **6132** also carry the I spreading code **6144**, the modulator **2604a** up-converts and spreads the filtered signal **6106** in an integrated manner during the sampling process. As such, the harmonic images in the harmonically rich signal **6108** are spread spectrum signals.

In step **6816**, the modulator **2604b** samples the Q filtered signal **6122** in a balanced and differential manner according to the control signals **6126** and **6128**, to generate a harmonically rich signal **6124**. The control signals **6126** and **6128** trigger the controlled switches in the modulator **2604b**, resulting in multiple harmonic images in the harmonically rich signal **6124**, where each image contains the Q baseband information. As with modulator **2604a**, the control signals **6126** and **6128** carry the Q spreading code **6142** so that the modulator **2604b** up-converts and spreads the filtered signal **6122** in an integrated manner during the sampling process. In other words, the harmonic images in the harmonically rich signal **6124** are also spread spectrum signals.

In step 6818, a 90 signal combiner 6146 combines the I harmonically rich signal 6108 and the Q harmonically rich signal 6124, to generate the IQ harmonically rich signal 6148. The IQ harmonically rich signal 6148 contains multiple harmonic images, where each image contains the spread I data and the spread Q data. The 90 degree combiner phase shifts the Q signal 6124 relative to the I signal 6108 so that no increase in spectrum width is needed for the IQ signal 6148, when compared to the I signal or the Q signal.

In step 6820, the optional bandpass filter 2606 selects the harmonic (or harmonics) of interest from the harmonically rich signal 6148, to generate signal 6114.

In step 6222, the optional amplifier 2608 amplifies the desired harmonic 6114 for transmission.

8.2 Integrated Up-Conversion and Spreading Using a Smoothing Varying Clock Signal

FIG. 54A illustrates a spread spectrum transmitter 5400 that is a second embodiment of balanced UFT modules that perform up-conversion and spreading simultaneously. More specifically, the spread spectrum transmitter 5400 does simultaneous up-conversion and spreading of an I data signal 5402a and a Q data signal 5402b to generate an IQ output signal 5428. Similar to modulator 6100, transmitter 5400 modulates the clock signal that controls the UFT modules with the spreading codes to spread the input I and Q signals during up-conversion. However, the transmitter 5400 modulates the clock signal by smoothly varying the instantaneous frequency or phase of a voltage controlled oscillator (VCO) with the spreading code. The transmitter 5400 is described in detail as follows with reference to a flowchart 6900 that is shown in FIGS. 69A and 69B.

In step 6901, the transmitter 5400 receives the I baseband signal 5402a and the Q baseband signal 5402b.

In step 6902, a code generator 5423 generates a spreading code 5422. In embodiments of the invention, the spreading code 5422 is a PN code or any other type of useful code for spread spectrum systems. Additionally, in embodiments of the invention, there are separate spreading codes for the I and Q channels.

In step 6904, a clock driver circuit 5421 generates a clock driver signal 5420 that is phase modulated according to a spreading code 5422. FIG. 54B illustrates the clock driver signal 5420 as series of pulses, where the instantaneous frequency (or phase) of the pulses is determined by the spreading code 5422, as shown. In embodiments of the invention, the phase of the pulses in the clock driver 5420 is varied smoothly in correlation with the spreading code 5422.

In step 6906, a voltage controlled oscillator 5418 generates a clock signal 5419 that has a frequency that varies according to a clock driver signal 5420. As mentioned above, the phase of the pulses in the clock driver 5420 is varied smoothly in correlation with the spreading code 5422 in embodiments of the invention. Since the clock driver 5420 controls the oscillator 5418, the frequency of the clock signal 5419 varies smoothly as a function of the PN code 5422. By smoothly varying the frequency of the clock signal 5419, the sidelobe growth in the spread spectrum images is minimized during the sampling process.

In step 6908, the pulse generator 2644 generates a control signal 5415 based on the clock signal 5419 that is similar to either one of the control signals 2623 or 2627 (in FIGS. 27A and 27B). The control signal 5415 carries the spreading code 5422 via the clock signal 5419. In embodiments of the invention, the pulse width ( $T_A$ ) of the control signal 5415 is established to enhance or optimize energy transfer to specific harmonics in the harmonically rich signal 5428 at the output. For the Q channel, a phase shifter 5414 shifts the phase of the

control signal 5415 by 90 degrees to implement the desired quadrature phase shift between the I and Q channels, resulting in a control signal 5413.

In step 6910, a low pass filter (LPF) 5406a filters the I data signal 5402a to remove any unwanted high frequency components, resulting in an I signal 5407a. Likewise, a LPF 5406b filters the Q data signal 5402b to remove any unwanted high frequency components, to generate the Q signal 5407b.

In step 6912, a UFT module 5408a samples the I data signal 5407a according to the control signal 5413 to generate a harmonically rich signal 5409a. The harmonically rich signal 5409a contains multiple spread spectrum harmonic images that repeat at harmonics of the sampling frequency. Similar to transmitter 5300, the harmonic images in signal 5409a carry the I baseband information, and are spread spectrum due to the spreading code on the control signal 5415.

In step 6914, a UFT module 5408b samples the Q data signal 5407b according to the control signal 5413 to generate a harmonically rich signal 5409b. The harmonically rich signal 5409b contains multiple spread spectrum harmonic images that repeat at harmonics of the sampling frequency. The harmonic images in signal 5409a carry the Q baseband information, and are spread spectrum due to the spreading code on the control signal 5413.

In step 6916, a signal combiner 5410 combines the harmonically rich signal 5409a with the harmonically rich signal 5409b to generate an IQ harmonically rich signal 5412. The harmonically rich signal 5412 carries multiple harmonic images, where each image carries the spread I data and the spread Q data.

In step 6918, the optional bandpass filter 5424 selects a harmonic (or harmonics) of interest for transmission, to generate the IQ output signal 5428.

FIG. 54C illustrates a transmitter 5430 that is similar to the transmitter 5400 except that the UFT modules are replaced by balanced UFT modulators 2604 that were described in FIG. 26. Also, the pulse generator is replaced by the control signal generator 2642 to generate the necessary control signals to operate the UFT modules in the balanced modulators. By replacing the UFT modules with balanced UFT modulators, sidelobe suppression can be improved.

#### 9.0 Shunt Receiver Embodiments Utilizing UFT Modules

In this section, example receiver embodiments are presented that utilize UFT modules in a differential and shunt configuration. More specifically, embodiments, according to the present invention, are provided for reducing or eliminating DC offset and/or reducing or eliminating circuit re-radiation in receivers, including I/Q modulation receivers and other modulation scheme receivers. These embodiments are described herein for purposes of illustration, and not limitation. The invention is not limited to these embodiments. Alternate embodiments (including equivalents, extensions, variations, deviations, etc., of the embodiments described herein) will be apparent to persons skilled in the relevant art(s) based on the teachings contained herein. The invention is intended and adapted to include such alternate embodiments.

#### 9.1 Example I/Q Modulation Receiver Embodiments

FIG. 70A illustrates an exemplary I/Q modulation receiver 7000, according to an embodiment of the present invention. I/Q modulation receiver 7000 has additional advantages of reducing or eliminating unwanted DC offsets and circuit re-radiation.

I/Q modulation receiver 7000 comprises a first UFD module 7002, a first optional filter 7004, a second UFD module 7006, a second optional filter 7008, a third UFD module 7010, a third optional filter 7012, a fourth UFD module 7014, a

fourth filter **7016**, an optional LNA **7018**, a first differential amplifier **7020**, a second differential amplifier **7022**, and an antenna **7072**.

I/Q modulation receiver **7000** receives, down-converts, and demodulates a I/Q modulated RF input signal **7082** to an I baseband output signal **7084**, and a Q baseband output signal **7086**. I/Q modulated RF input signal **7082** comprises a first information signal and a second information signal that are I/Q modulated onto an RF carrier signal. I baseband output signal **7084** comprises the first baseband information signal. Q baseband output signal **7086** comprises the second baseband information signal.

Antenna **7072** receives I/Q modulated RF input signal **7082**. I/Q modulated RF input signal **7082** is output by antenna **7072** and received by optional LNA **7018**. When present, LNA **7018** amplifies I/Q modulated RF input signal **7082**, and outputs amplified I/Q signal **7088**.

First UFD module **7002** receives amplified I/Q signal **7088**. First UFD module **7002** down-converts the I-phase signal portion of amplified input I/Q signal **7088** according to an I control signal **7090**. First UFD module **7002** outputs an I output signal **7098**.

In an embodiment, first UFD module **7002** comprises a first storage module **7024**, a first UFT module **7026**, and a first voltage reference **7028**. In an embodiment, a switch contained within first UFT module **7026** opens and closes as a function of I control signal **7090**. As a result of the opening and closing of this switch, which respectively couples and de-couples first storage module **7024** to and from first voltage reference **7028**, a down-converted signal, referred to as I output signal **7098**, results. First voltage reference **7028** may be any reference voltage, and is preferably ground. I output signal **7098** is stored by first storage module **7024**.

In an embodiment, first storage module **7024** comprises a first capacitor **7074**. In addition to storing I output signal **7098**, first capacitor **7074** reduces or prevents a DC offset voltage resulting from charge injection from appearing on I output signal **7098**.

I output signal **7098** is received by optional first filter **7004**. When present, first filter **7004** is in some embodiments a high pass filter to at least filter I output signal **7098** to remove any carrier signal “bleed through”. In a preferred embodiment, when present, first filter **7004** comprises a first resistor **7030**, a first filter capacitor **7032**, and a first filter voltage reference **7034**. Preferably, first resistor **7030** is coupled between I output signal **7098** and a filtered I output signal **7007**, and first filter capacitor **7032** is coupled between filtered I output signal **7007** and first filter voltage reference **7034**. Alternately, first filter **7004** may comprise any other applicable filter configuration as would be understood by persons skilled in the relevant art(s). First filter **7004** outputs filtered I output signal **7007**.

Second UFD module **7006** receives amplified I/Q signal **7088**. Second UFD module **7006** down-converts the inverted I-phase signal portion of amplified input I/Q signal **7088** according to an inverted I control signal **7092**. Second UFD module **7006** outputs an inverted I output signal **7001**.

In an embodiment, second UFD module **7006** comprises a second storage module **7036**, a second UFT module **7038**, and a second voltage reference **7040**. In an embodiment, a switch contained within second UFT module **7038** opens and closes as a function of inverted I control signal **7092**. As a result of the opening and closing of this switch, which respectively couples and de-couples second storage module **7036** to and from second voltage reference **7040**, a down-converted signal, referred to as inverted I output signal **7001**, results. Second voltage reference **7040** may be any reference voltage,

and is preferably ground. Inverted I output signal **7001** is stored by second storage module **7036**.

In an embodiment, second storage module **7036** comprises a second capacitor **7076**. In addition to storing inverted I output signal **7001**, second capacitor **7076** reduces or prevents a DC offset voltage resulting from charge injection from appearing on inverted I output signal **7001**.

Inverted I output signal **7001** is received by optional second filter **7008**. When present, second filter **7008** is a high pass filter to at least filter inverted I output signal **7001** to remove any carrier signal “bleed through”. In a preferred embodiment, when present, second filter **7008** comprises a second resistor **7042**, a second filter capacitor **7044**, and a second filter voltage reference **7046**. Preferably, second resistor **7042** is coupled between inverted I output signal **7001** and a filtered inverted I output signal **7009**, and second filter capacitor **7044** is coupled between filtered inverted I output signal **7009** and second filter voltage reference **7046**. Alternately, second filter **7008** may comprise any other applicable filter configuration as would be understood by persons skilled in the relevant art(s). Second filter **7008** outputs filtered inverted I output signal **7009**.

First differential amplifier **7020** receives filtered I output signal **7007** at its non-inverting input and receives filtered inverted I output signal **7009** at its inverting input. First differential amplifier **7020** subtracts filtered inverted I output signal **7009** from filtered I output signal **7007**, amplifies the result, and outputs I baseband output signal **7084**. Because filtered inverted I output signal **7009** is substantially equal to an inverted version of filtered I output signal **7007**, I baseband output signal **7084** is substantially equal to filtered I output signal **7009**, with its amplitude doubled. Furthermore, filtered I output signal **7007** and filtered inverted I output signal **7009** may comprise substantially equal noise and DC offset contributions from prior down-conversion circuitry, including first UFD module **7002** and second UFD module **7006**, respectively. When first differential amplifier **7020** subtracts filtered inverted I output signal **7009** from filtered I output signal **7007**, these noise and DC offset contributions substantially cancel each other.

Third UFD module **7010** receives amplified I/Q signal **7088**. Third UFD module **7010** down-converts the Q-phase signal portion of amplified input I/Q signal **7088** according to an Q control signal **7094**. Third UFD module **7010** outputs an Q output signal **7003**.

In an embodiment, third UFD module **7010** comprises a third storage module **7048**, a third UFT module **7050**, and a third voltage reference **7052**. In an embodiment, a switch contained within third UFT module **7050** opens and closes as a function of Q control signal **7094**. As a result of the opening and closing of this switch, which respectively couples and de-couples third storage module **7048** to and from third voltage reference **7052**, a down-converted signal, referred to as Q output signal **7003**, results. Third voltage reference **7052** may be any reference voltage, and is preferably ground. Q output signal **7003** is stored by third storage module **7048**.

In an embodiment, third storage module **7048** comprises a third capacitor **7078**. In addition to storing Q output signal **7003**, third capacitor **7078** reduces or prevents a DC offset voltage resulting from charge injection from appearing on Q output signal **7003**.

Q output signal **7003** is received by optional third filter **7012**. When present, in an embodiment, third filter **7012** is a high pass filter to at least filter Q output signal **7003** to remove any carrier signal “bleed through”. In an embodiment, when present, third filter **7012** comprises a third resistor **7054**, a third filter capacitor **7056**, and a third filter voltage reference

7058. Preferably, third resistor 7054 is coupled between Q output signal 7003 and a filtered Q output signal 7011, and third filter capacitor 7056 is coupled between filtered Q output signal 7011 and third filter voltage reference 7058. Alternately, third filter 7012 may comprise any other applicable filter configuration as would be understood by persons skilled in the relevant art(s). Third filter 7012 outputs filtered Q output signal 7011.

Fourth UFD module 7014 receives amplified I/Q signal 7088. Fourth UFD module 7014 down-converts the inverted Q-phase signal portion of amplified input I/Q signal 7088 according to an inverted Q control signal 7096. Fourth UFD module 7014 outputs an inverted Q output signal 7005.

In an embodiment, fourth UFD module 7014 comprises a fourth storage module 7060, a fourth UFT module 7062, and a fourth voltage reference 7064. In an embodiment, a switch contained within fourth UFT module 7062 opens and closes as a function of inverted Q control signal 7096. As a result of the opening and closing of this switch, which respectively couples and de-couples fourth storage module 7060 to and from fourth voltage reference 7064, a down-converted signal, referred to as inverted Q output signal 7005, results. Fourth voltage reference 7064 may be any reference voltage, and is preferably ground. Inverted Q output signal 7005 is stored by fourth storage module 7060.

In an embodiment, fourth storage module 7060 comprises a fourth capacitor 7080. In addition to storing inverted Q output signal 7005, fourth capacitor 7080 reduces or prevents a DC offset voltage resulting from charge injection from appearing on inverted Q output signal 7005.

Inverted Q output signal 7005 is received by optional fourth filter 7016. When present, fourth filter 7016 is a high pass filter to at least filter inverted Q output signal 7005 to remove any carrier signal "bleed through". In a preferred embodiment, when present, fourth filter 7016 comprises a fourth resistor 7066, a fourth filter capacitor 7068, and a fourth filter voltage reference 7070. Preferably, fourth resistor 7066 is coupled between inverted Q output signal 7005 and a filtered inverted Q output signal 7013, and fourth filter capacitor 7068 is coupled between filtered inverted Q output signal 7013 and fourth filter voltage reference 7070. Alternately, fourth filter 7016 may comprise any other applicable filter configuration as would be understood by persons skilled in the relevant art(s). Fourth filter 7016 outputs filtered inverted Q output signal 7013.

Second differential amplifier 7022 receives filtered Q output signal 7011 at its non-inverting input and receives filtered inverted Q output signal 7013 at its inverting input. Second differential amplifier 7022 subtracts filtered inverted Q output signal 7013 from filtered Q output signal 7011, amplifies the result, and outputs Q baseband output signal 7086. Because filtered inverted Q output signal 7013 is substantially equal to an inverted version of filtered Q output signal 7011, Q baseband output signal 7086 is substantially equal to filtered Q output signal 7013, with its amplitude doubled. Furthermore, filtered Q output signal 7011 and filtered inverted Q output signal 7013 may comprise substantially equal noise and DC offset contributions of the same polarity from prior down-conversion circuitry, including third UFD module 7010 and fourth UFD module 7014, respectively. When second differential amplifier 7022 subtracts filtered inverted Q output signal 7013 from filtered Q output signal 7011, these noise and DC offset contributions substantially cancel each other.

Additional embodiments relating to addressing DC offset and re-radiation concerns, applicable to the present invention, are described in co-pending patent application No., "DC Offset, Re-radiation, and I/Q Solutions Using Universal Fre-

quency Translation Technology," Ser. No. 11/059,536, which is herein incorporated by reference in its entirety.

9.1.1 Example I/Q Modulation Control Signal Generator Embodiments

FIG. 70B illustrates an exemplary block diagram for I/Q modulation control signal generator 7023, according to an embodiment of the present invention. I/Q modulation control signal generator 7023 generates I control signal 7090, inverted I control signal 7092, Q control signal 7094, and inverted Q control signal 7096 used by I/Q modulation receiver 7000 of FIG. 70A. I control signal 7090 and inverted I control signal 7092 operate to down-convert the I-phase portion of an input I/Q modulated RF signal. Q control signal 7094 and inverted Q control signal 7096 act to down-convert the Q-phase portion of the input I/Q modulated RF signal. Furthermore, I/Q modulation control signal generator 7023 has the advantage of generating control signals in a manner such that resulting collective circuit re-radiation is radiated at one or more frequencies outside of the frequency range of interest. For instance, potential circuit re-radiation is radiated at a frequency substantially greater than that of the input RF carrier signal frequency.

I/Q modulation control signal generator 7023 comprises a local oscillator 7025, a first divide-by-two module 7027, a 180 degree phase shifter 7029, a second divide-by-two module 7031, a first pulse generator 7033, a second pulse generator 7035, a third pulse generator 7037, and a fourth pulse generator 7039.

Local oscillator 7025 outputs an oscillating signal 7015. FIG. 70C shows an exemplary oscillating signal 7015.

First divide-by-two module 7027 receives oscillating signal 7015, divides oscillating signal 7015 by two, and outputs a half frequency LO signal 7017 and a half frequency inverted LO signal 7041. FIG. 70C shows an exemplary half frequency LO signal 7017. Half frequency inverted LO signal 7041 is an inverted version of half frequency LO signal 7017. First divide-by-two module 7027 may be implemented in circuit logic, hardware, software, or any combination thereof, as would be known by persons skilled in the relevant art(s).

180 degree phase shifter 7029 receives oscillating signal 7015, shifts the phase of oscillating signal 7015 by 180 degrees, and outputs phase shifted LO signal 7019. 180 degree phase shifter 7029 may be implemented in circuit logic, hardware, software, or any combination thereof, as would be known by persons skilled in the relevant art(s). In alternative embodiments, other amounts of phase shift may be used.

Second divide-by two module 7031 receives phase shifted LO signal 7019, divides phase shifted LO signal 7019 by two, and outputs a half frequency phase shifted LO signal 7021 and a half frequency inverted phase shifted LO signal 7043. FIG. 70C shows an exemplary half frequency phase shifted LO signal 7021. Half frequency inverted phase shifted LO signal 7043 is an inverted version of half frequency phase shifted LO signal 7021. Second divide-by-two module 7031 may be implemented in circuit logic, hardware, software, or any combination thereof, as would be known by persons skilled in the relevant art(s).

First pulse generator 7033 receives half frequency LO signal 7017, generates an output pulse whenever a rising edge is received on half frequency LO signal 7017, and outputs I control signal 7090. FIG. 70C shows an exemplary I control signal 7090.

Second pulse generator 7035 receives half frequency inverted LO signal 7041, generates an output pulse whenever a rising edge is received on half frequency inverted LO signal

**7041**, and outputs inverted I control signal **7092**. FIG. **70C** shows an exemplary inverted I control signal **7092**.

Third pulse generator **7037** receives half frequency phase shifted LO signal **7021**, generates an output pulse whenever a rising edge is received on half frequency phase shifted LO signal **7021**, and outputs Q control signal **7094**. FIG. **70C** shows an exemplary Q control signal **7094**.

Fourth pulse generator **7039** receives half frequency inverted phase shifted LO signal **7043**, generates an output pulse whenever a rising edge is received on half frequency inverted phase shifted LO signal **7043**, and outputs inverted Q control signal **7096**. FIG. **70C** shows an exemplary inverted Q control signal **7096**.

In an embodiment, control signals **7090**, **7021**, **7041** and **7043** include pulses having a width equal to one-half of a period of I/Q modulated RF input signal **7082**. The invention, however, is not limited to these pulse widths, and control signals **7090**, **7021**, **7041**, and **7043** may comprise pulse widths of any fraction of, or multiple and fraction of, a period of I/Q modulated RF input signal **7082**.

First, second, third, and fourth pulse generators **7033**, **7035**, **7037**, and **7039** may be implemented in circuit logic, hardware, software, or any combination thereof, as would be known by persons skilled in the relevant art(s).

As shown in FIG. **70C**, in an embodiment, control signals **7090**, **7021**, **7041**, and **7043** comprise pulses that are non-overlapping in other embodiments the pulses may overlap. Furthermore, in this example, pulses appear on these signals in the following order: I control signal **7090**, Q control signal **7094**, inverted I control signal **7092**, and inverted Q control signal **7096**. Potential circuit re-radiation from I/Q modulation receiver **7000** may comprise frequency components from a combination of these control signals.

For example, FIG. **70D** shows an overlay of pulses from I control signal **7090**, Q control signal **7094**, inverted I control signal **7092**, and inverted Q control signal **7096**. When pulses from these control signals leak through first, second, third, and/or fourth UFD modules **7002**, **7006**, **7010**, and **7014** to antenna **7072** (shown in FIG. **70A**), they may be radiated from I/Q modulation receiver **7000**, with a combined waveform that appears to have a primary frequency equal to four times the frequency of any single one of control signals **7090**, **7021**, **7041**, and **7043**. FIG. **70** shows an example combined control signal **7045**.

FIG. **70D** also shows an example I/Q modulation RF input signal **7082** overlaid upon control signals **7090**, **7094**, **7092**, and **7096**. As shown in FIG. **70D**, pulses on I control signal **7090** overlay and act to down-convert a positive I-phase portion of I/Q modulation RE input signal **7082**. Pulses on inverted I control signal **7092** overlay and act to down-convert a negative I-phase portion of I/Q modulation RF input signal **7082**. Pulses on Q control signal **7094** overlay and act to down-convert a rising Q-phase portion of I/Q modulation RF input signal **7082**. Pulses on inverted Q control signal **7096** overlay and act to down-convert a falling Q-phase portion of I/Q modulation RF input signal **7082**.

As FIG. **70D** further shows in this example, the frequency ratio between the combination of control signals **7090**, **7021**, **7041**, and **7043** and I/Q modulation RF input signal **7082** is approximately 4:3. Because the frequency of the potentially re-radiated signal, i.e., combined control signal **7045**, is substantially different from that of the signal being down-converted, i.e., I/Q modulation RF input signal **7082**, it does not interfere with signal down-conversion as it is out of the frequency band of interest, and hence may be filtered out. In this manner, I/Q modulation receiver **7000** reduces problems due to circuit re-radiation. As will be understood by persons

skilled in the relevant art(s) from the teachings herein, frequency ratios other than 4:3 may be implemented to achieve similar reduction of problems of circuit re-radiation.

It should be understood that the above control signal generator circuit example is provided for illustrative purposes only. The invention is not limited to these embodiments. Alternative embodiments (including equivalents, extensions, variations, deviations, etc., of the embodiments described herein) for I/Q modulation control signal generator **7023** will be apparent to persons skilled in the relevant art(s) from the teachings herein, and are within the scope of the present invention.

Additional embodiments relating to addressing DC offset and re-radiation concerns, applicable to the present invention, are described in co-pending patent application titled "DC Offset, Re-radiation, and I/Q Solutions Using Universal Frequency Translation Technology," which is herein incorporated by reference in its entirety.

9.1.2 Detailed Example I/Q Modulation Receiver Embodiment with Exemplary Waveforms

FIG. **70E** illustrates a more detailed example circuit implementation of I/Q modulation receiver **7000**, according to an embodiment of the present invention. FIGS. **70E-P** show example waveforms related to an example implementation of I/Q modulation receiver **7000** of FIG. **70E**.

FIGS. **70F** and **70G** show first and second input data signals **7047** and **7049** to be I/Q modulated with a RF carrier signal frequency as the I-phase and Q-phase information signals, respectively.

FIGS. **70I** and **70J** show the signals of FIGS. **70F** and **70G** after modulation with a RF carrier signal frequency, respectively, as I-modulated signal **7051** and Q-modulated signal **7053**.

FIG. **70H** shows an I/Q modulation RF input signal **7082** formed from I-modulated signal **7051** and Q-modulated signal **7053** of FIGS. **70I** and **70J**, respectively.

FIG. **70O** shows an overlaid view of filtered I output signal **7007** and filtered inverted I output signal **7009**.

FIG. **70P** shows an overlaid view of filtered Q output signal **7011** and filtered inverted Q output signal **7013**.

FIGS. **70K** and **70L** show I baseband output signal **7084** and Q baseband output signal **7086**, respectfully. A data transition **7055** is indicated in both I baseband output signal **7084** and Q baseband output signal **7086**. The corresponding data transition **7055** is indicated in I-modulated signal **7051** of FIG. **70I**, Q-modulated signal **7053** of FIG. **70J**, and I/Q modulation RF input signal **7082** of FIG. **70H**.

FIGS. **70M** and **70N** show I baseband output signal **7084** and Q baseband output signal **7086** over a wider time interval.

9.2 Example Single Channel Receiver Embodiment

FIG. **70Q** illustrates an example single channel receiver **7091**, corresponding to either the I or Q channel of I/Q modulation receiver **7000**, according to an embodiment of the present invention. Single channel receiver **7091** can down-convert an input RF signal **7097** modulated according to AM, PM, FM, and other modulation schemes. Refer to section 7.4.1 above for further description on the operation of single channel receiver **7091**.

9.3 Alternative Example I/Q Modulation Receiver Embodiment

FIG. **70R** illustrates an exemplary I/Q modulation receiver **7089**, according to an embodiment of the present invention. I/Q modulation receiver **7089** receives, down-converts, and demodulates an I/Q modulated RF input signal **7082** to an I baseband output signal **7084**, and a Q baseband output signal **7086**. I/Q modulation receiver **7089** has additional advantages of reducing or eliminating unwanted DC offsets and

circuit re-radiation, in a similar fashion to that of I/Q modulation receiver **7000** described above.

#### 10. Shunt Transceiver Embodiments using UFT Modules

In this section, example transceiver embodiments are presented that utilize UFT modules in a shunt configuration for balanced up-conversion and balanced down-conversion.

More specifically, a signal channel transceiver embodiment is presented that incorporates the balanced transmitter **5600** (FIG. **56A**) and the receiver **7091** (FIG. **70Q**). Additionally, an IQ transceiver embodiment is presented that incorporate balanced IQ transmitter **5700** (FIG. **57**) and IQ receiver **7000** (FIG. **70A**).

These transceiver embodiments incorporate the advantages described above for the balanced transmitter **5600** and the balanced receiver **7091**. More specifically, during up-conversion, an input baseband signal is up-converted in a balanced and differential fashion, so as to minimize carrier insertion and unwanted spectral growth. Additionally, during down-conversion, an input RF input signal is down-converted so that DC offset and re-radiation is reduced or eliminated. Additionally, since both transmitter and receiver utilize UFT modules for frequency translation, integration and cost saving can be realized.

These embodiments are described herein for purposes of illustration, and not limitation. The invention is not limited to these embodiments. Alternate embodiments (including equivalents, extensions, variations, deviations, etc., of the embodiments described herein) will be apparent to persons skilled in the relevant art(s) based on the teachings contained herein. The invention is intended and adapted to include such alternate embodiments.

FIG. **71** illustrates a transceiver **7100** according to embodiments of the present invention. Transceiver **7100** includes the single channel receiver **7091**, the balanced transmitter **5600**, a diplexer **7108**, and an antenna **7112**. Transceiver **7100** up-converts a baseband input signal **7110** using the balanced transmitter **5600** resulting in an output RF signal **7106** that is radiated by the antenna **7112**. Additionally, the transceiver **7100** also down-converts a received RF input signal **7104** using the receiver **7091** to output baseband signal **7102**. The diplexer **7108** separates the transmit signal **7106** from the receive signal **7104** so that the same antenna **7112** can be used for both transmit and receive operations. The operation of transmitter **5600** is described above in section 7.1.3, to which the reader is referred for greater detail.

During up-conversion, the transmitter **5600** shunts the input baseband signal **7110** to ground in a differential and balanced fashion according to the control signals **2623** and **2627**, resulting in the harmonically rich signal **7114**. The harmonically rich signal **7114** includes multiple harmonic images that repeat at harmonics of the sampling frequency of the control signals, where each harmonic image contains the necessary amplitude, frequency, and phase information to reconstruct the baseband signal **7110**. The optional filter **2606** can be included to select a desired harmonic from the harmonically rich signal **7114**. The optional amplifier **2608** can be included to amplify the desired harmonic resulting in the output RF signal **7106**, which is transmitted by antenna **7112** after the diplexer **7108**. A detailed description of the transmitter **5600** is included in section 7.1.3, to which the reader is referred for further details.

During down-conversion, the receiver **7091** alternately shunts the received RF signal **7104** to ground according to control signals **7093** and **7095**, resulting in the down-converted output signal **7102**. A detailed description of receiver **7091** is included in sections 9.1 and 9.2, to which the reader is referred for further details.

FIG. **72** illustrates IQ transceiver **7200** according to embodiments of the present invention. IQ transceiver **7200** includes the IQ receiver **7000**, the IQ transmitter **5700**, a diplexer **7214**, and an antenna **7216**. Transceiver **7200** up-converts an I baseband signal **7206** and a Q baseband signal **7208** using the IQ transmitter **5700** (FIG. **57**) to generate an IQ RF output signal **7212**. A detailed description of the IQ transmitter **5700** is included in section 7.2.2, to which the reader is referred for further details. Additionally, the transceiver **7200** also down-converts a received RF signal **7210** using the IQ Receiver **7000**, resulting in I baseband output signal **7202** and a Q baseband output signal **7204**. A detailed description of the IQ receiver **7000** is included in section 9.1, to which the reader is referred for further details.

#### 11. Conclusion

Example implementations of the methods, systems and components of the invention have been described herein. As noted elsewhere, these example implementations have been described for illustrative purposes only, and are not limiting. Other implementation embodiments are possible and covered by the invention, such as but not limited to software and software/hardware implementations of the systems and components of the invention. Such implementation embodiments will be apparent to persons skilled in the relevant art(s) based on the teachings contained herein.

While various application embodiments of the present invention have been described above, it should be understood that they have been presented by way of example only, and not limitation. Thus, the breadth and scope of the present invention should not be limited by any of the above-described exemplary embodiments.

What is claimed is:

**1.** A method for up-converting a baseband signal, to a higher frequency signal comprising the steps for:

- (1) receiving the baseband signal;
- (2) using first and second control signals, each having a single fundamental frequency, the first and second control signals being phase shifted from one another, and each control signal having variable pulse widths, to differentially sample the baseband signal to generate both I and Q harmonically rich signals each containing a plurality of harmonic images, the I and Q harmonically rich signals each being a function of information of the baseband signal;
- (3) controlling the amplitude of each harmonically rich signal by adjusting the variable pulse widths of the first and second control signals;
- (4) combining corresponding portions of harmonic images for the harmonically rich signals and then filtering the combined result for a selected harmonic frequency to obtain an up-converted signal;
- (5) selecting a desired harmonic from said harmonic images; and
- (6) transmitting said desired harmonic over a communications medium.

**2.** The method of claim **1**, further comprising the step of: minimizing DC offset voltages during step (2), and thereby minimizing carrier insertion in said harmonic images.

**3.** The method of claim **1**, wherein the controlling step further comprises maintaining a reference voltage between said differential samples.

**4.** The method of claim **1**, wherein step (2) comprises the steps of: (a) converting said baseband signal into a differential baseband signal having a first differential baseband component and a second differential baseband component; (b) sampling said first differential baseband component according to said first control signal to generate the first harmonically rich

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signal, and sampling said second differential baseband component according to said second control signal to generate the second harmonically rich signal, wherein said second control signal is phase shifted relative to said first control signal as measured by a master clock signal; and (c) combining said first harmonically rich signal and said second harmonically rich signal to generate said harmonic images.

5. The method of claim 4, further comprising the step of: (d) adding a reference voltage to said first differential baseband component and said second differential baseband component prior to step (b), and thereby minimizing any DC offset voltages during sampling of said first differential baseband component and said second differential baseband component.

6. The method of claim 4, wherein said step (b) of sampling comprises the steps of: (i) generating said first control signal comprising a first plurality of pulses and said second control signal comprising a second plurality of pulses; and (ii) operating a first switch according to said first control signal to periodically sample said first differential baseband component, and operating a second switch according to said second control signal to periodically sample said second differential baseband signal.

7. The method of claim 6, wherein said step (i) comprises the step of controlling pulse widths of said first control signal and said second control signal by a specified amount to control a time interval that said first switch and said second switch is closed in step (ii), and thereby controlling energy transfer to said desired harmonic image.

8. The method of claim 7, wherein said step of controlling pulse widths comprises the step of controlling pulse widths for said first and second control signals to a non-zero fraction of a period of a desired harmonic of interest.

9. The method of claim 7, wherein said step of controlling pulse widths comprises the step of controlling pulse widths for said first and second control signals to approximately one-half of a period of a desired harmonic of interest.

10. The method of claim 7, wherein said step of controlling pulse widths comprises the step of controlling pulse widths for said first and second control signals to approximately one-fourth of a period of a desired harmonic of interest.

11. The method of claim 7, wherein said step of controlling pulse widths comprises the step of controlling pulse widths for said first and second control signals to approximately one-tenth to one-fourth of a period of a desired harmonic of interest.

12. The method of claim 1, wherein said first control signal and said second control signal have a period of  $T_s$  so that said harmonic images repeat at  $1/T_s$  in frequency, and wherein said second control signal is phase-shifted relative to said first control signal by approximately 180 degrees.

13. The method of claim 1, wherein said first control signal and said second control signal have a period of  $T_s$  so that said harmonic images repeat at  $1/T_s$  in frequency, and wherein said second control signal is phase-shifted relative to said first control signal by approximately 90 degrees.

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14. The method of claim 1, wherein said pulse widths of said first control signal and said second control signal are a non-zero fraction of a period of a desired harmonic of interest.

15. The method of claim 1, wherein said pulse widths of said first control signal and said second control signal are approximately one-half of a period of a desired harmonic of interest.

16. The method of claim 1, wherein said pulse widths of said first control signal and said second control signal are approximately one-half of a period of a desired harmonic of interest.

17. The method of claim 1, wherein said plurality of harmonic images have an amplitude that is proportional to the following equation:  $\text{Amp} = [4 \sin(n \cdot \pi \cdot T_A / T_s) \sin(n \cdot \pi \cdot 2 / n \cdot \pi)]$  where:  $T_s$  = period of said first and second control signals,  $T_A$  = pulse width of said first and second control signals,  $n$  = harmonic number of said harmonic image whose amplitude is determined.

18. The method of claim 1, wherein said harmonic images have an amplitude that is based on  $n \cdot (T_A / T_s)$ , where  $T_s$  is a period of said first and second control signals,  $T_A$  is a pulse width of pulses in said first and second control signals, and  $n$  is a harmonic number of said harmonic image.

19. The method of claim 1, wherein control signals having substantially shorter pulse widths shift an increased amount of energy into higher frequency harmonics.

20. The method of claim 1, wherein control signals having relatively longer pulse widths shift an increased amount of energy into lower frequency harmonics.

21. The method of claim 1, wherein the information of the baseband signal includes amplitude, or phase or frequency or any combination thereof.

22. A differential frequency up-conversion module for up-converting a baseband signal to a higher frequency signal, comprising:

- an input terminal for receiving at least one baseband signal;
- first and second switching devices for receiving, respectively, first and second control signals having a single fundamental frequency, the first and second control signals being phase shifted from one another, and each control signal having variable pulse widths to differentially sample the baseband signal in order to generate I and Q harmonically rich signals, each containing a plurality of harmonic images, the I and Q harmonically rich signals each being representative of the baseband signal, and having an amplitude adjusted based on the pulse widths of the first and second control signals;
- a combiner that combines corresponding portions of particular ones of the harmonic images for the harmonically rich signals; and
- a filter that filters results of the combiner for a selected harmonic frequency to obtain an up-converted signal.

23. The differential frequency up-conversion module as claimed in claim 22, wherein the combiner provides a direct connection between the first and second switching devices.

24. The differential frequency up-conversion module as claimed in claim 22, wherein the combiner is a wire.

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