US009923518B2

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

(10) **Patent No.: US 9,923,518 B2**
(45) **Date of Patent: Mar. 20, 2018**

(54) **PHASE-SWITCHED IMPEDANCE MODULATION AMPLIFIER**

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(73) Assignee: **Massachusetts Institute of Technology**, Cambridge, MA (US)

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

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(Continued)

(51) **Int. Cl.**

H03F 1/02	(2006.01)
H03F 3/193	(2006.01)
H03F 3/217	(2006.01)
H03F 1/56	(2006.01)
H03F 3/19	(2006.01)
H03F 3/21	(2006.01)

(Continued)

(52) **U.S. Cl.**

CPC ***H03F 1/0205*** (2013.01); ***H03F 1/56*** (2013.01); ***H03F 1/565*** (2013.01); ***H03F 3/19*** (2013.01); ***H03F 3/193*** (2013.01); ***H03F 3/21*** (2013.01); ***H03F 3/2171*** (2013.01); ***H03F 3/2176*** (2013.01); ***H03F 3/245*** (2013.01); ***H03H 7/38*** (2013.01); *H03F 2200/301* (2013.01); *H03F 2200/387* (2013.01); *H03F 2200/391* (2013.01); *H03F 2200/451* (2013.01)

(58) **Field of Classification Search**

USPC 330/305, 151, 136, 277

See application file for complete search history.

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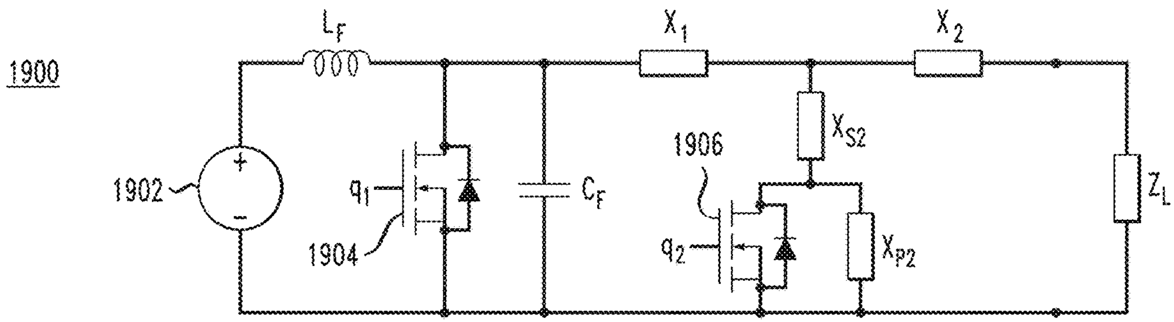
(57) **ABSTRACT**

Described is a phase-switched phase-switched, impedance modulation amplifier (PSIM). The PSIM has an input that can be coupled to a source and an output that can be coupled to a load. The PSIM includes one or more phase-switched reactive elements and a controller. The controller provides a control signal to each of the one or more phase-switched reactive elements. In response to one or more control signals provided thereto, each phase-switched reactive element provides a corresponding selected reactance value such that an impedance presented to an input or output port of an RF amplifier may be varied.

30 Claims, 17 Drawing Sheets



Related U.S. Application Data

(60) Provisional application No. 62/094,144, filed on Dec. 19, 2014.

(51) **Int. Cl.**

H03F 3/24	(2006.01)
H03H 7/38	(2006.01)

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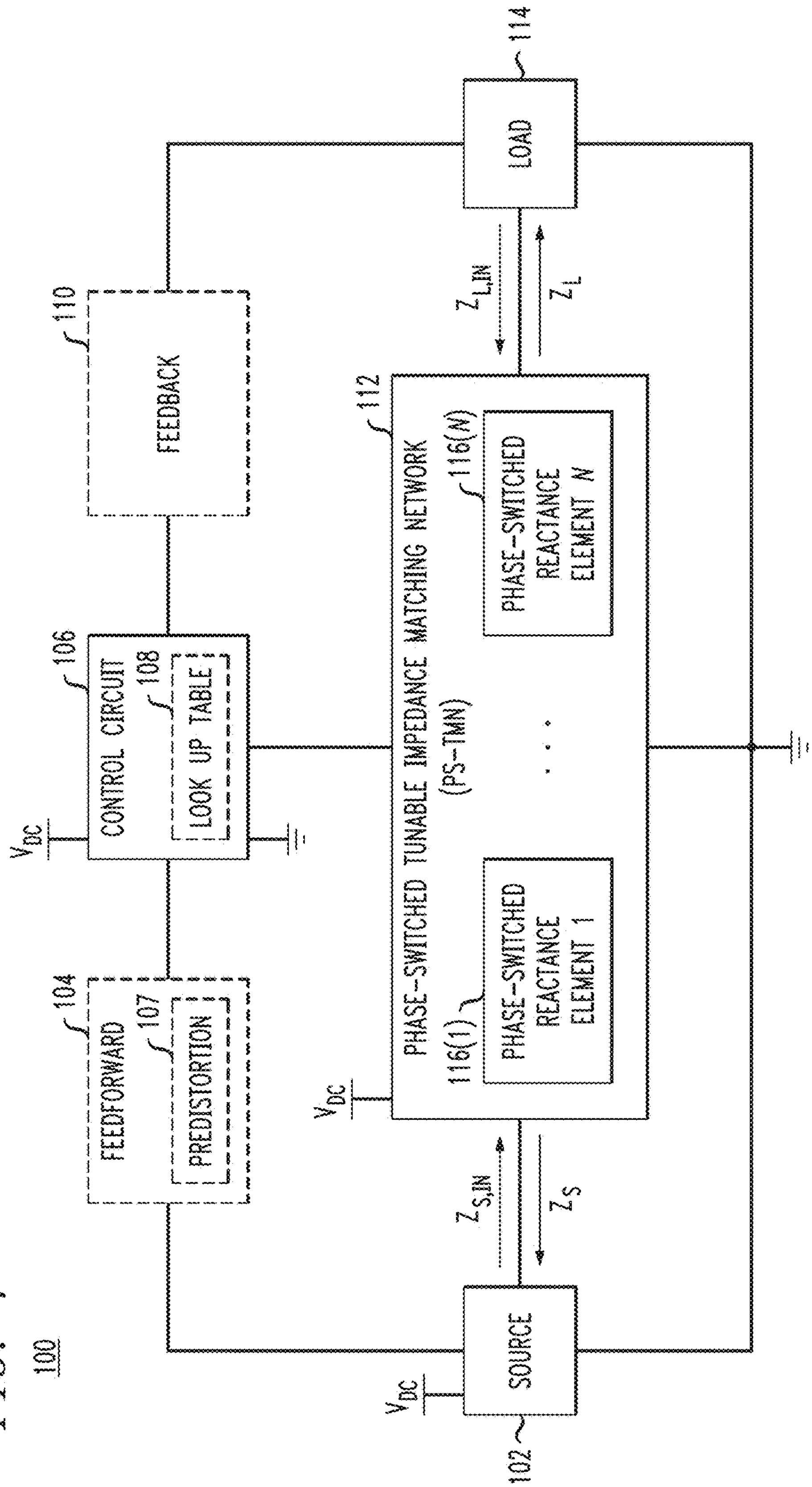

SOURCE
102
V_{DC}
$Z_{S,IN}$
Z_S
FEEDFORWARD
104
PREDISTORTION
107
CONTROL CIRCUIT
106
LOOK UP TABLE
108
V_{DC}
FEEDBACK
110
PHASE-SWITCHED TUNABLE IMPEDANCE MATCHING NETWORK (PS-TMN)
112
V_{DC}
PHASE-SWITCHED REACTANCE ELEMENT 1
116(1)
. . .
PHASE-SWITCHED REACTANCE ELEMENT N
116(N)
$Z_{L,IN}$
Z_L
LOAD
114

FIG. 1
100

FIG. 2

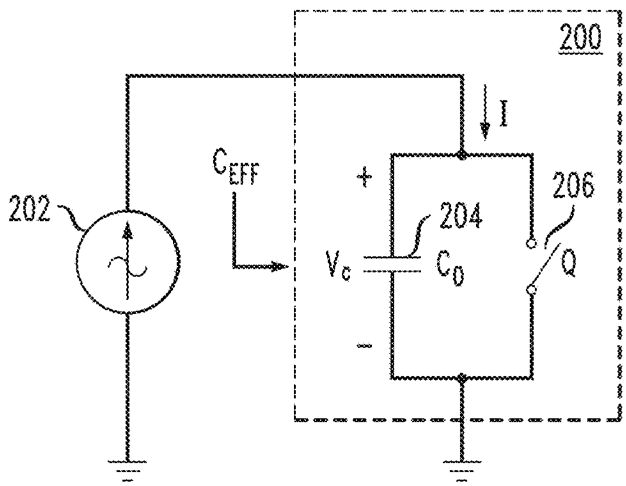

200
I
C_{EFF}
+
206
202
204
V_c
C_0
Q
−

FIG. 3

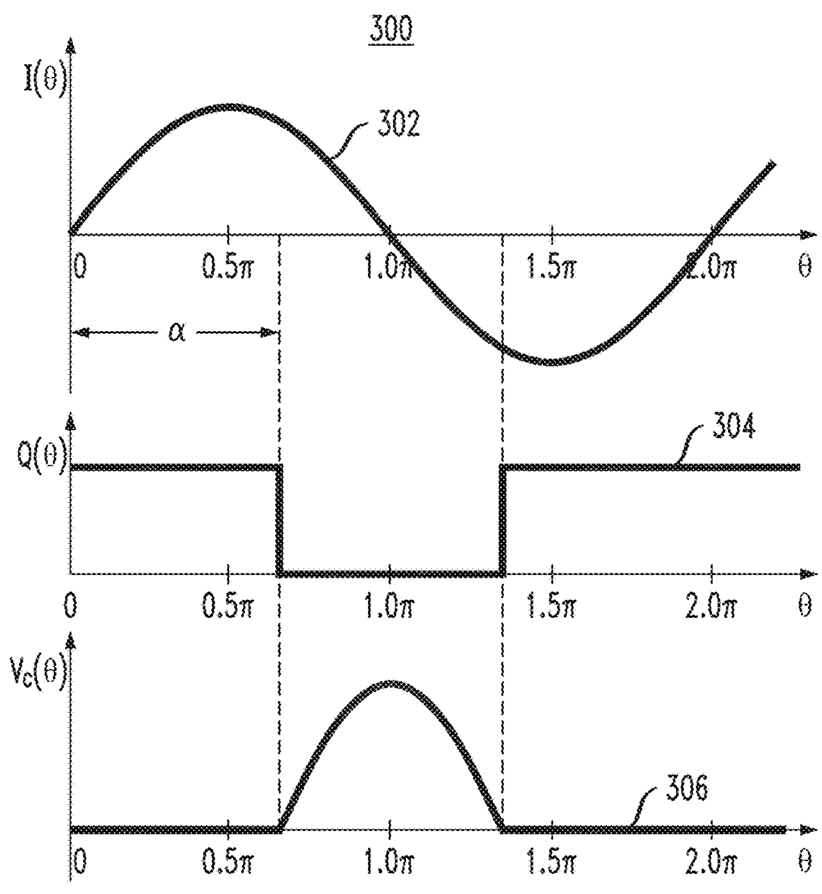

300
I(θ)
302
0
0.5π
1.0π
1.5π
2.0π
θ
α
Q(θ)
304
V_c(θ)
306

FIG. 4

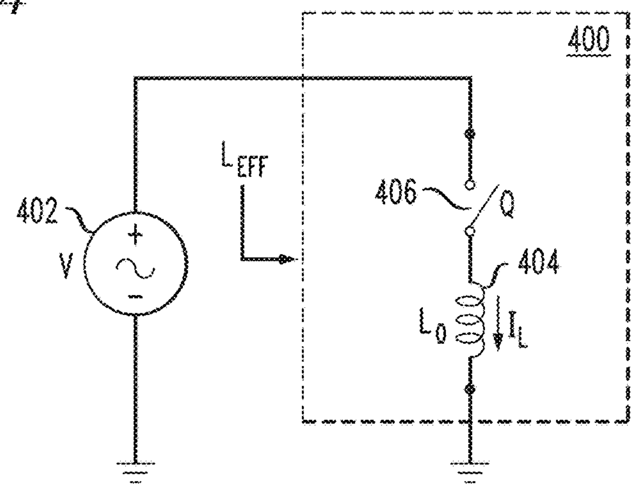

400
L_{EFF}
402
V
406
Q
404
L_0
I_L

FIG. 5

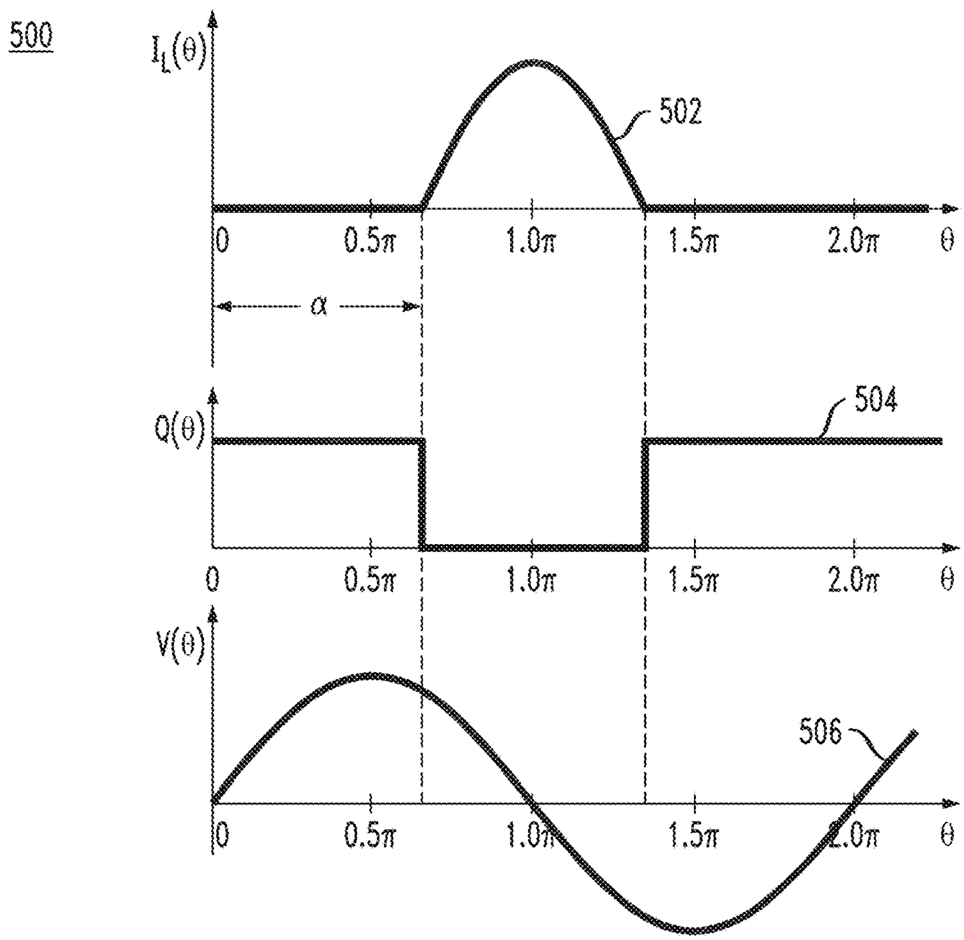

500
$I_L(\theta)$
502
0
0.5π
1.0π
1.5π
2.0π
θ
α
$Q(\theta)$
504
$V(\theta)$
506

FIG. 6

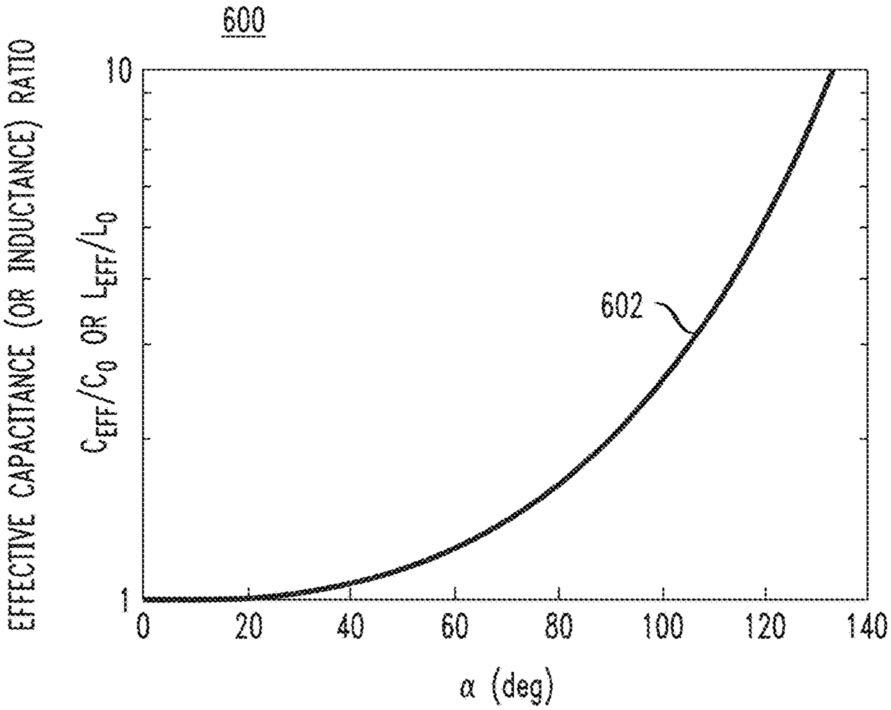

600
EFFECTIVE CAPACITANCE (OR INDUCTANCE) RATIO
C_{EFF}/C_0 OR L_{EFF}/L_0
10
1
0
20
40
60
80
100
120
140
602
α (deg)

FIG. 7

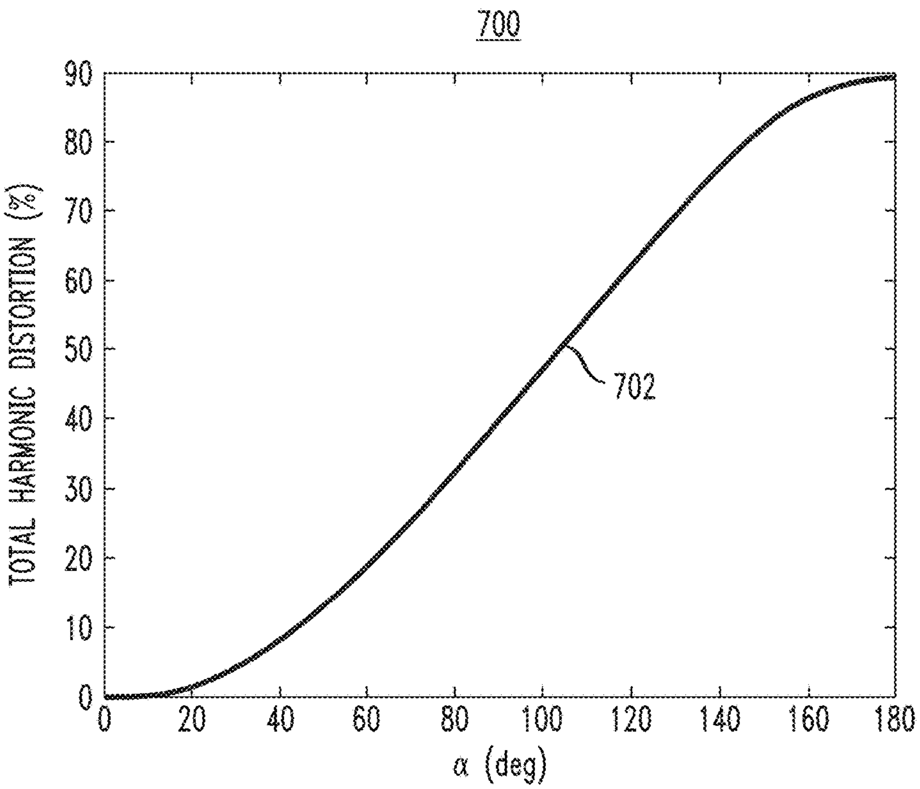

700
TOTAL HARMONIC DISTORTION (%)
90
80
70
60
50
40
30
20
10
0
0
20
40
60
80
100
120
140
160
180
702
α (deg)

FIG. 8

800

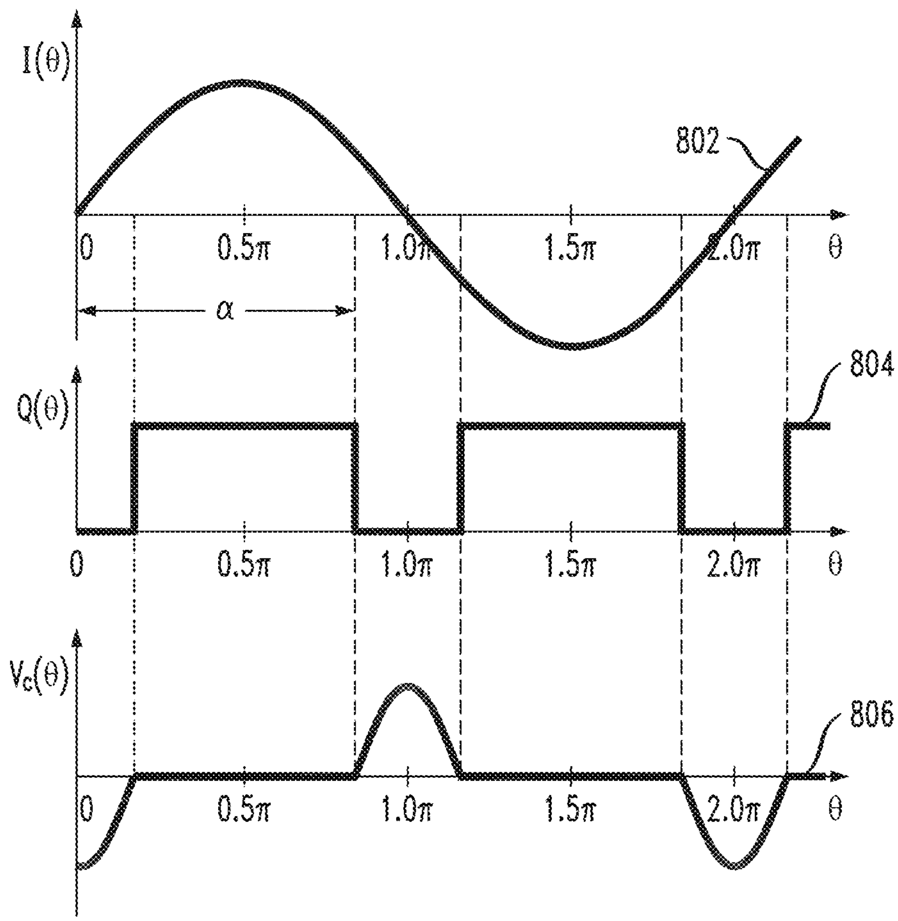

I(θ)
802
0
0.5π
1.0π
1.5π
2.0π
θ
α
Q(θ)
804
0
0.5π
1.0π
1.5π
2.0π
θ
Vc(θ)
806
0
0.5π
1.0π
1.5π
2.0π
θ

FIG. 9

900

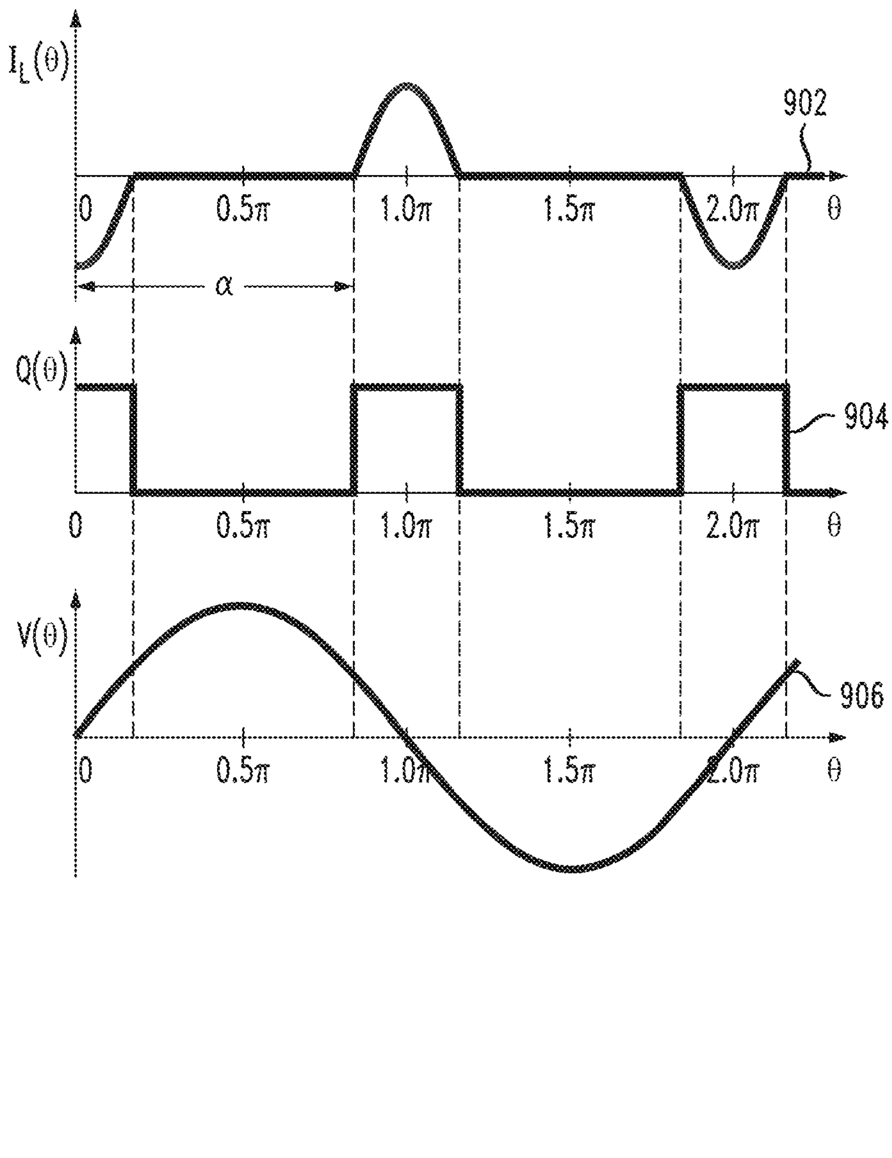


FIG. 10D

1008

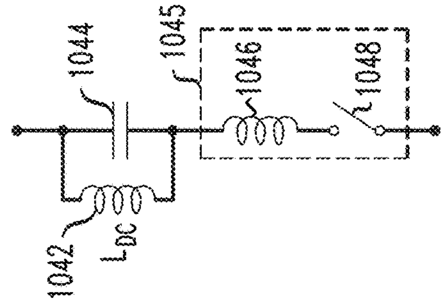


FIG. 10C

1006

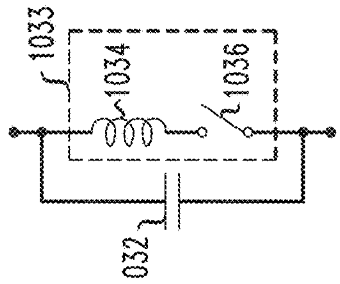


FIG. 10B

1004

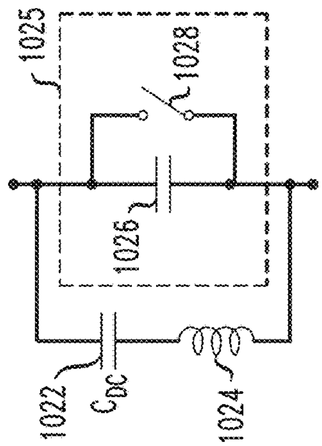
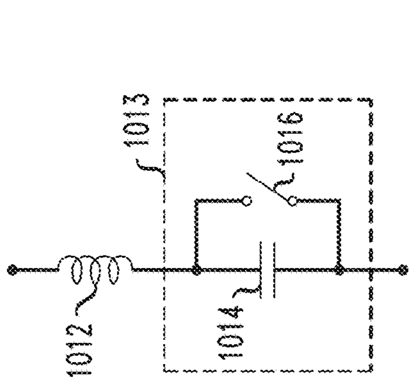
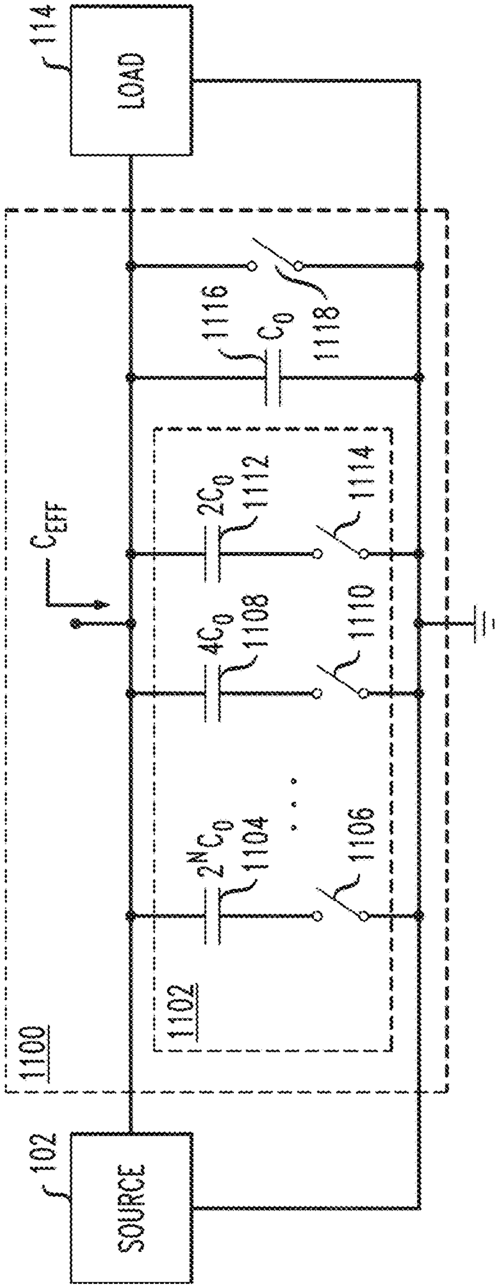


FIG. 10A

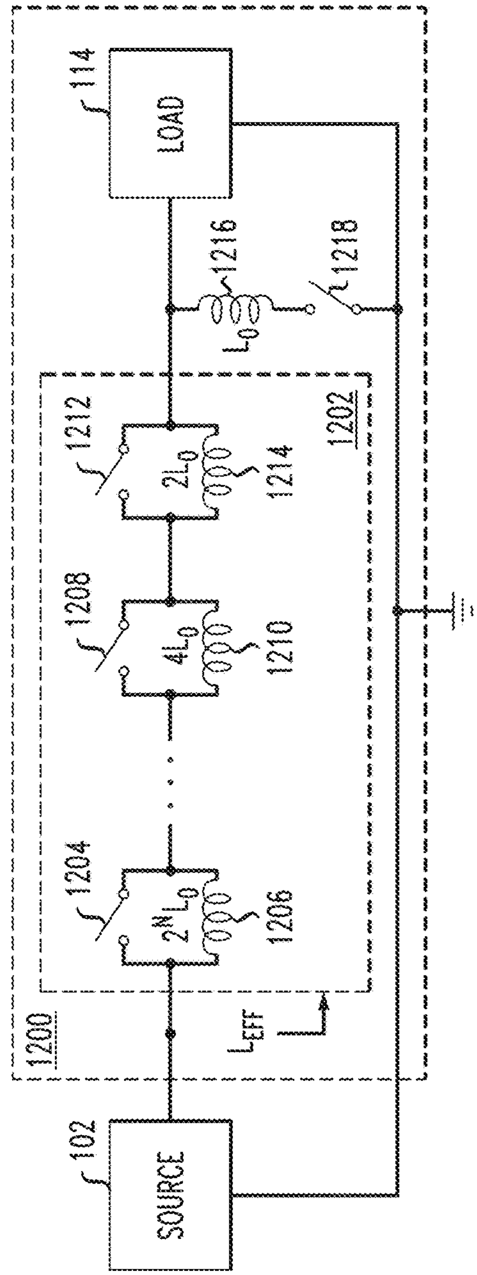
1002





SOURCE
102
LOAD
114
1100
1102
C_EFF
2^N C_0
1104
1106
4C_0
1108
1110
2C_0
1112
1114
C_0
1116
1118

FIG. 11



SOURCE
102
LOAD
114
1200
1202
L_EFF
2^N L_0
1204
1206
4L_0
1208
1210
2L_0
1212
1214
L_0
1216
1218

FIG. 12

FIG. 13

1300

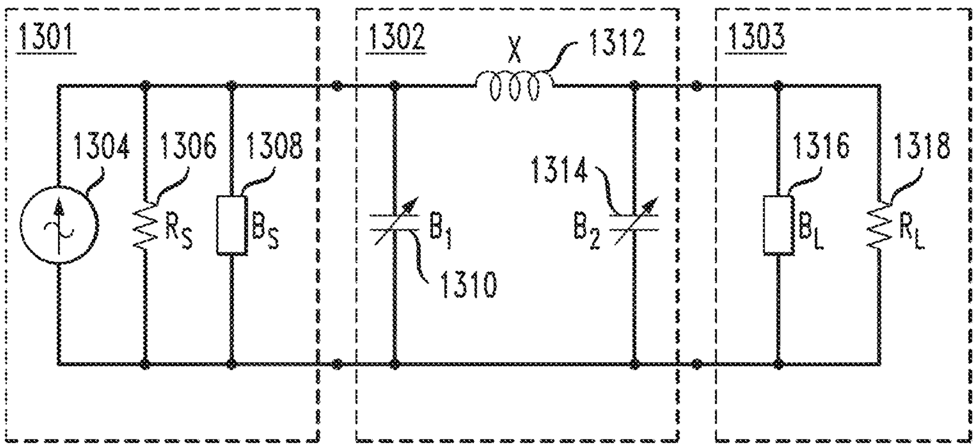


FIG. 14

1400

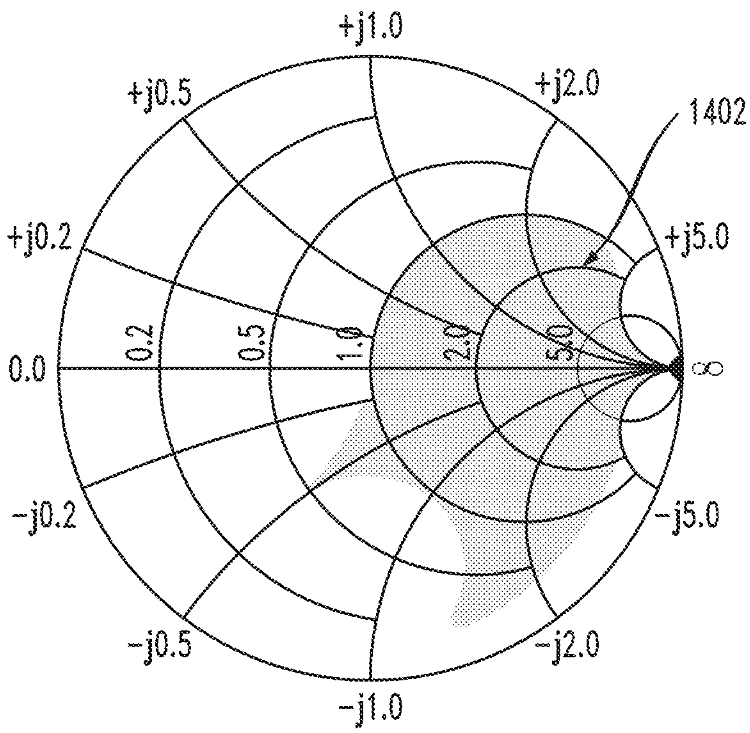


FIG. 15

1500

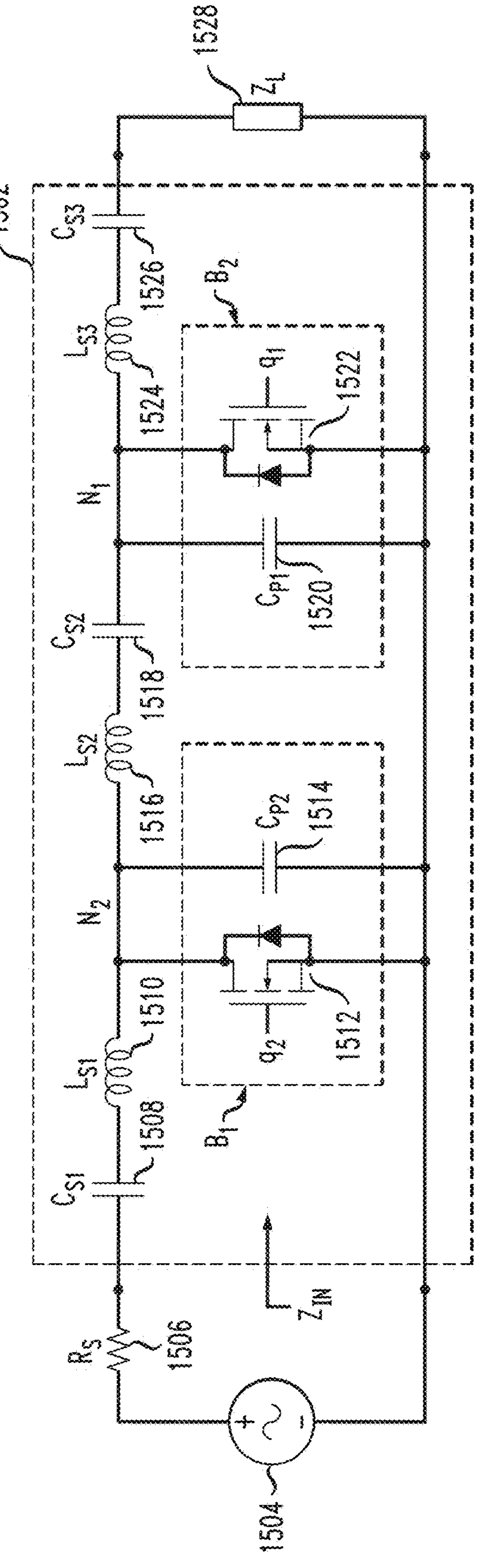

1528
Z_L
1502
C_{S3}
1526
L_{S3}
1524
B_2
q_1
1522
N_1
C_{P1}
1520
C_{S2}
1518
L_{S2}
1516
C_{P2}
1514
N_2
1510
L_{S1}
1508
q_2
1512
B_1
C_{S1}
Z_{IN}
R_S
1506
+
-
1504

FIG. 16

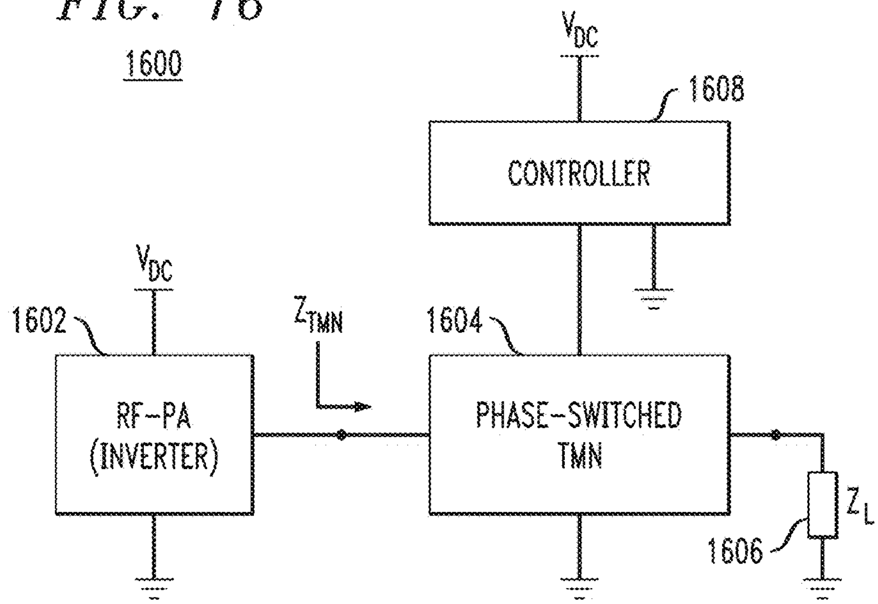
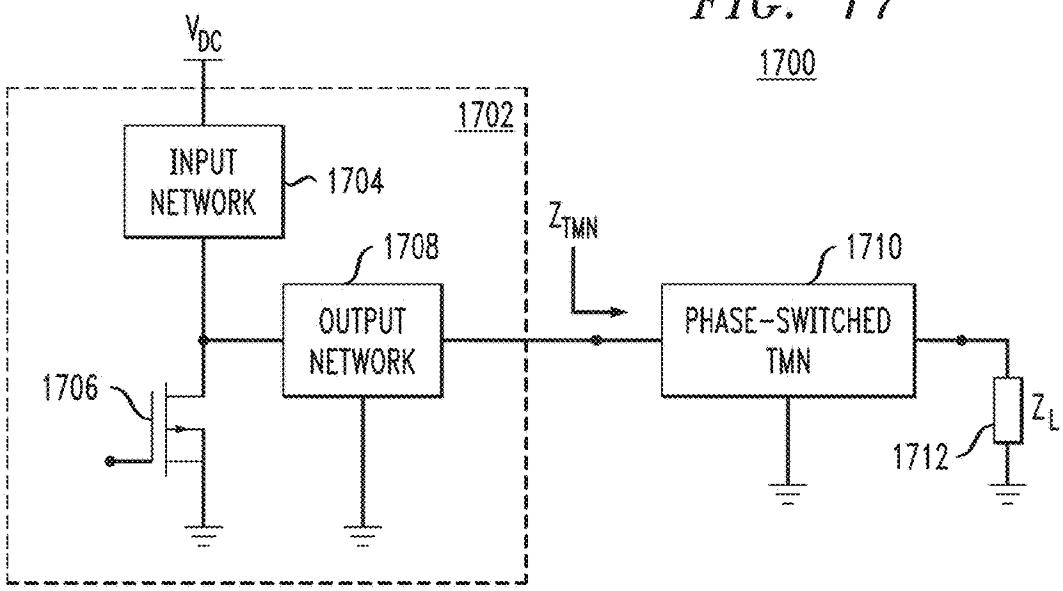

1600
V_{DC}
1608
CONTROLLER
V_{DC}
1602
Z_{TMN}
1604
RF-PA
(INVERTER)
PHASE-SWITCHED
TMN
Z_L
1606

FIG. 17


1700
V_{DC}
1702
INPUT
NETWORK
1704
Z_{TMN}
1708
1710
OUTPUT
NETWORK
PHASE-SWITCHED
TMN
1706
Z_L
1712

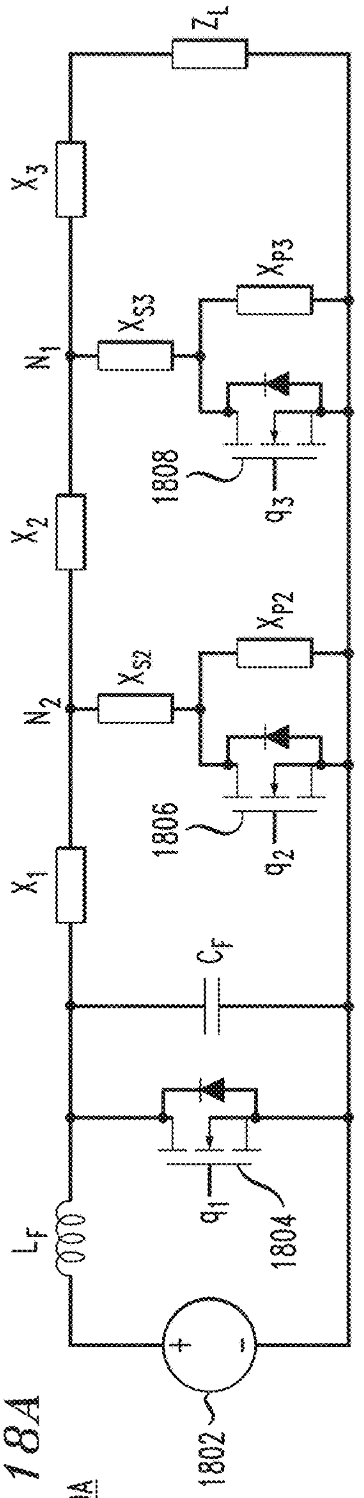


FIG. 18A
1800A

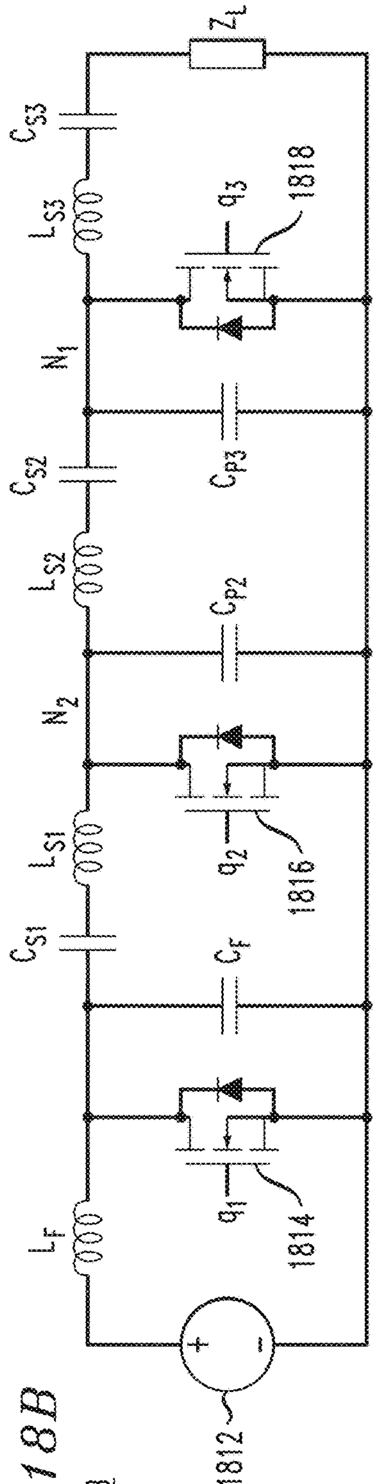


FIG. 18B
1800B

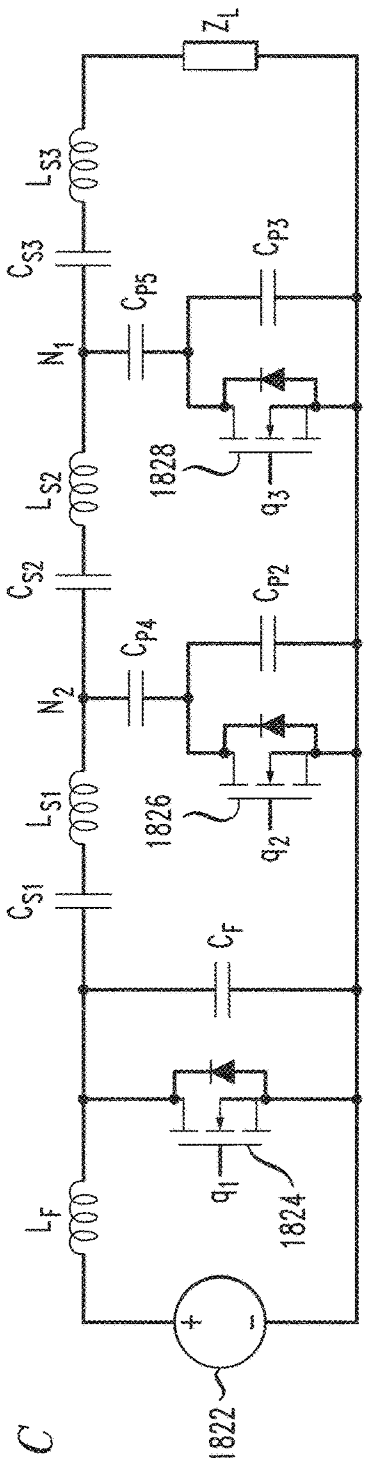


FIG. 18C
1800C

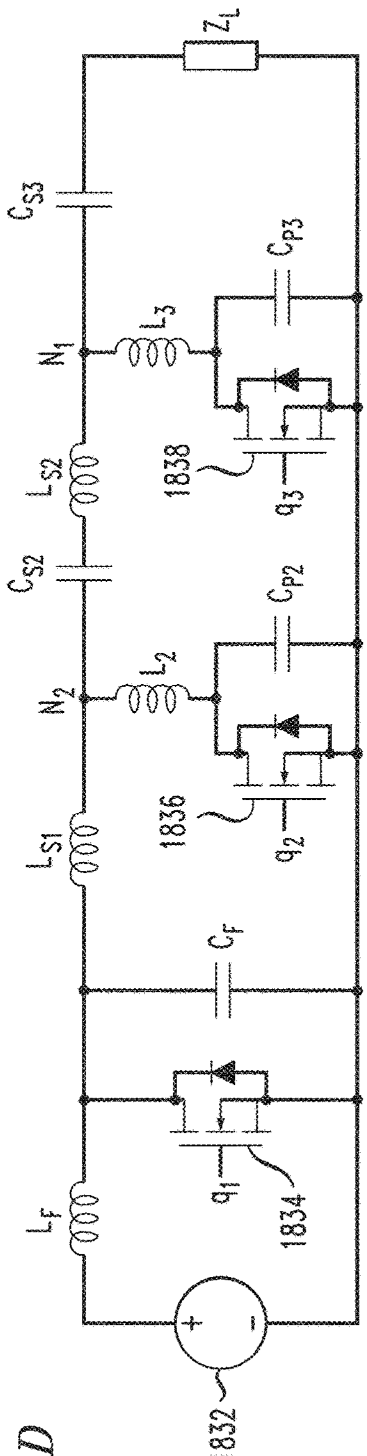


FIG. 18D
1800D

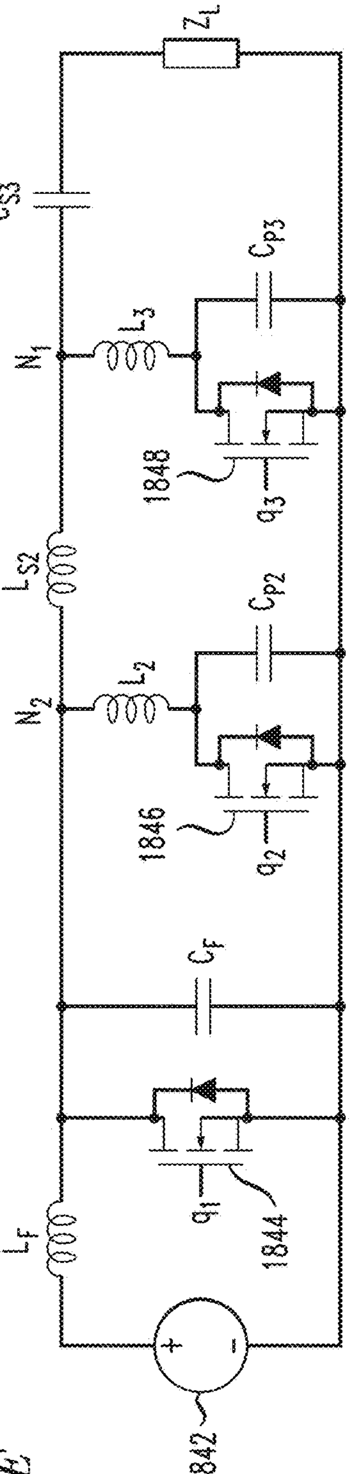


FIG. 18E
1800E

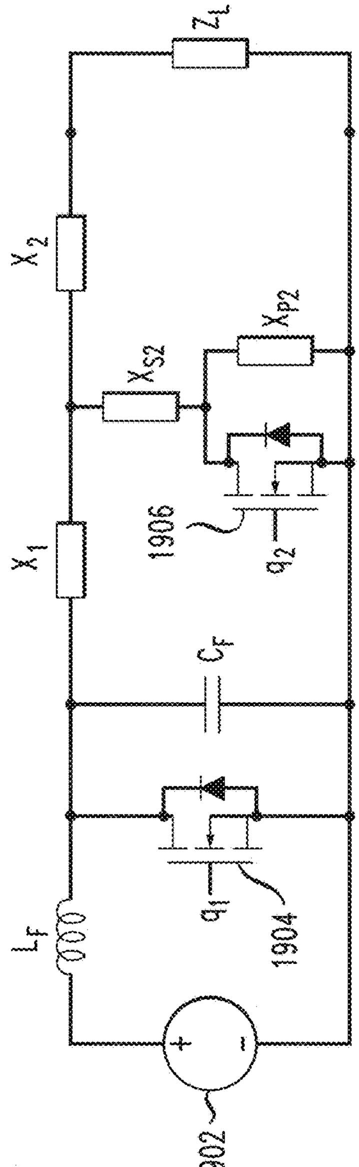


FIG. 19
1900

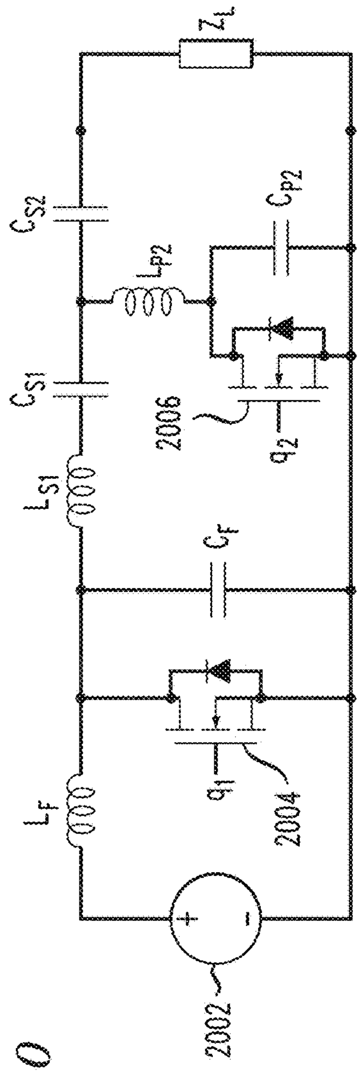


FIG. 20
2000

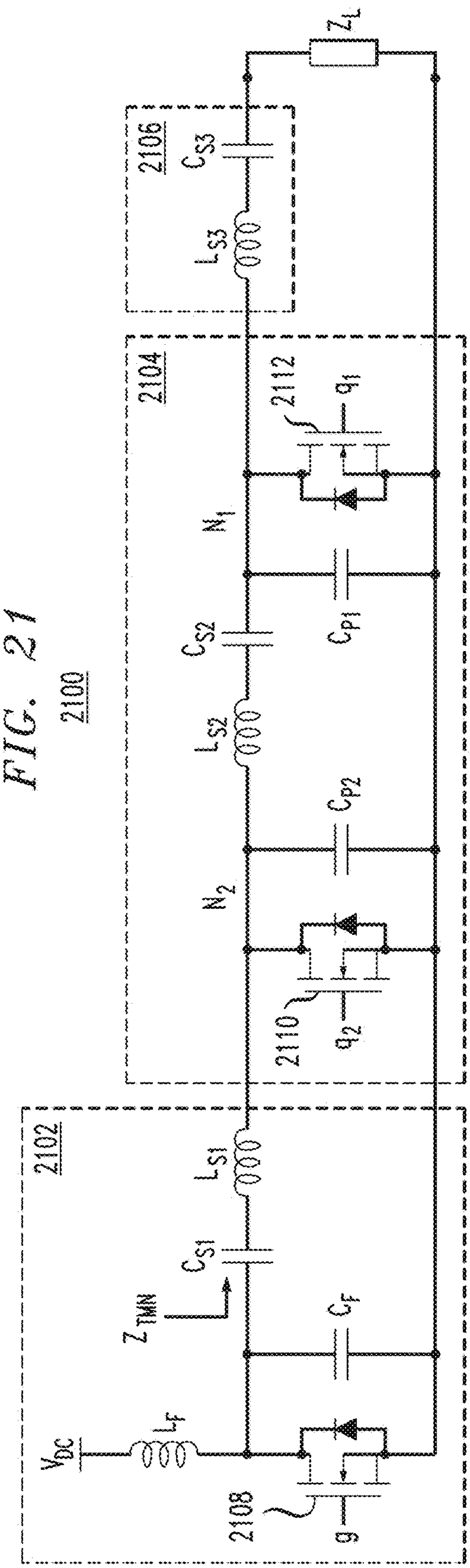


FIG. 21
2100

FIG. 22

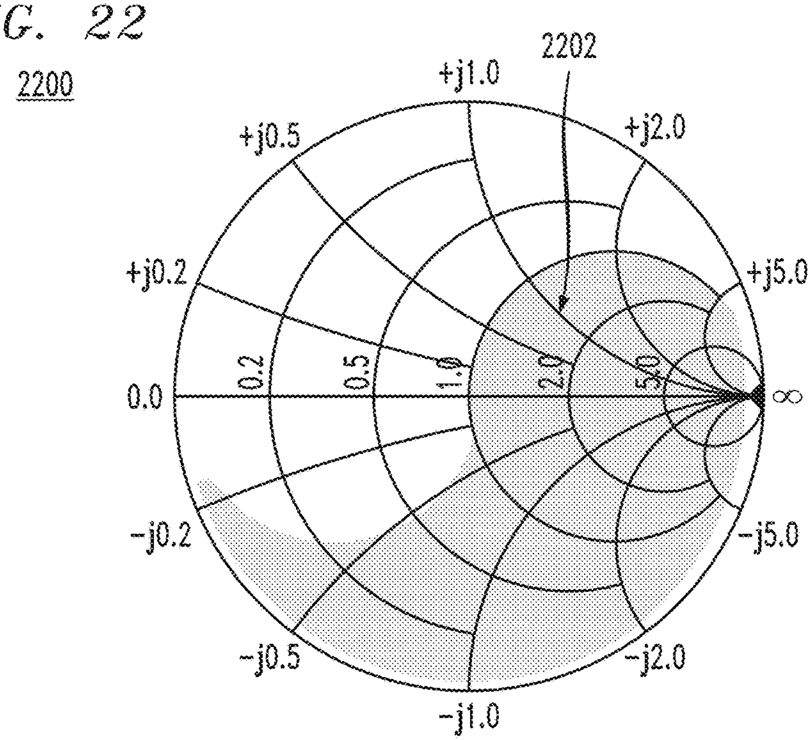

2200
2202
+j1.0
+j0.5
+j2.0
+j0.2
+j5.0
0.0
0.2
0.5
1.0
2.0
5.0
∞
−j0.2
−j5.0
−j0.5
−j2.0
−j1.0

FIG. 23

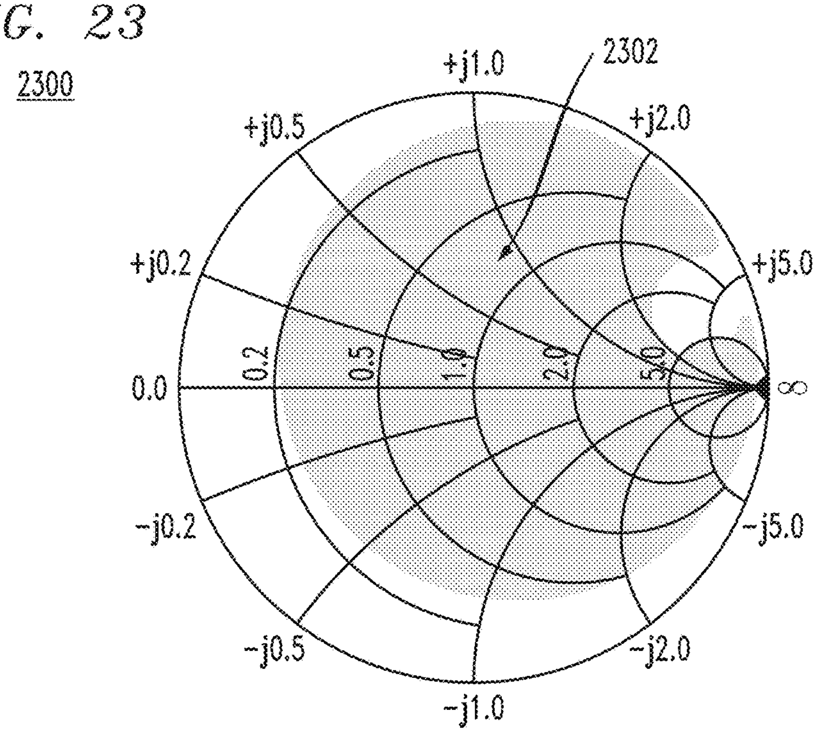
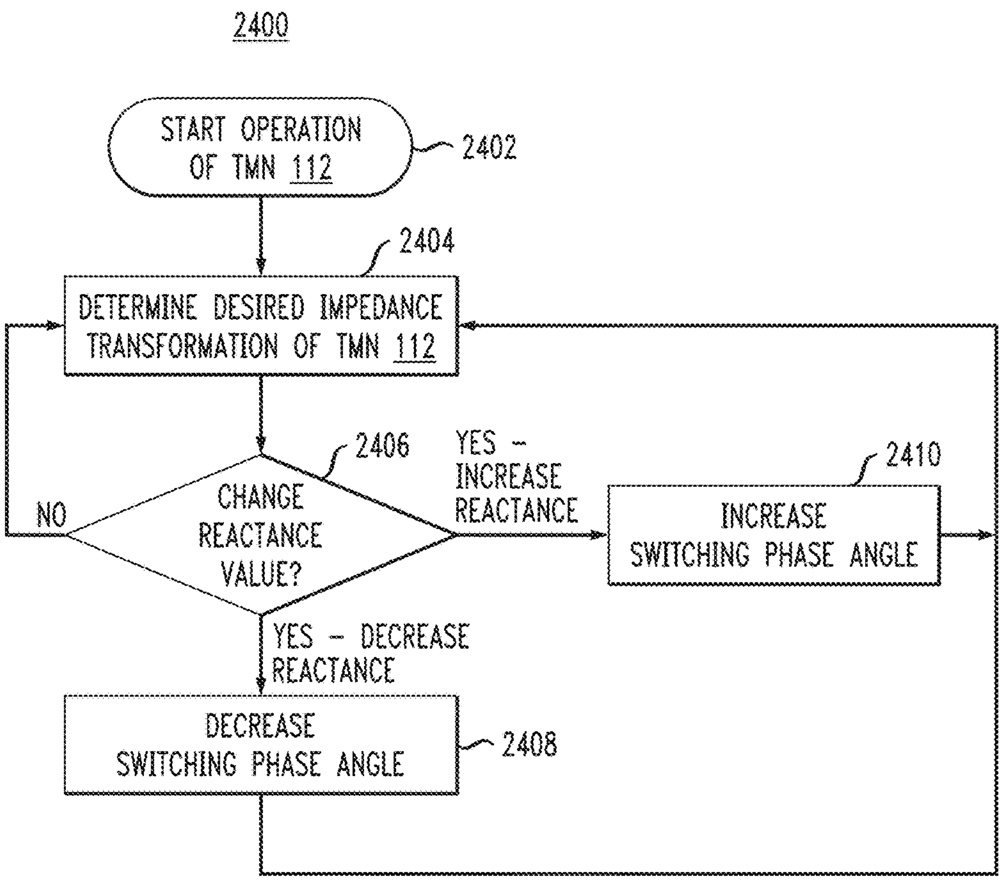

2300
2302
+j1.0
+j0.5
+j2.0
+j0.2
+j5.0
0.0
0.2
0.5
1.0
2.0
5.0
∞
−j0.2
−j5.0
−j0.5
−j2.0
−j1.0

FIG. 24


2400
START OPERATION OF TMN 112
2402
2404
DETERMINE DESIRED IMPEDANCE TRANSFORMATION OF TMN 112
2406
CHANGE REACTANCE VALUE?
NO
YES - INCREASE REACTANCE
2410
INCREASE SWITCHING PHASE ANGLE
YES - DECREASE REACTANCE
DECREASE SWITCHING PHASE ANGLE
2408

PHASE-SWITCHED IMPEDANCE MODULATION AMPLIFIER

CROSS-REFERENCE TO RELATED APPLICATIONS

This application claims the benefit under 35 U.S.C. § 119(e) of U.S. provisional application No. 62/094,144, filed on Dec. 19, 2014, which application is hereby incorporated herein by reference in its entirety. This application is a continuation of and claims the benefit under 35 U.S.C. § 120 of U.S. application Ser. No. 14/974,563, filed on Dec. 18, 2015, which application is hereby incorporated herein by reference in its entirety.

BACKGROUND

Impedance matching networks are commonly used for maximizing power transfer within many radio frequency (RF) and microwave systems. For example, in RF transmitters, impedance matching networks might be used to provide an impedance match from an output impedance of an RF power amplifier (PA) to an impedance of an RF load (e.g., an antenna). Such impedance matching increases the transmitted power, reduces power loss and reduces or eliminates the need for additional circuit elements (e.g., isolators, etc.).

One class of impedance matching networks is referred to as tunable impedance matching networks (TMNs), sometimes called automatic antenna tuning units. Conventional TMNs might be implemented as single-element or lumped-element reactive networks where at least one of the reactive elements are variable (e.g., tunable) components such that the impedance of the variable components at a particular frequency, or over a range of frequencies, can be modified. The reactive elements within a TMN might be arranged in circuit topologies such as a ladder-network, an L-network, a T-network, or a Pi-network.

Conventional TMNs can be classified as either analog (continuously adjustable) or digital (adjustable among a set of discrete values). Analog TMNs utilize variable reactance elements having reactance values (at some frequency or over a range of frequencies) that can be tuned in a continuous manner by adjusting bias conditions. Digital TMNs implement the variable reactive elements as digitally switched arrays of static reactance elements. This approach allows adjustment of the impedance of the reactance values in finite and discrete steps.

Analog TMNs are often implemented using varactor diodes (or varactor diode circuits) or micro-electromechanical systems (MEMS) varactors. Although analog TMNs allow fast and accurate impedance matching over a wide range of impedances, relatively high bias voltages are required to operate at high power levels.

Digital TMNs are often implemented using CMOS switches, MEMS switches, PIN diodes or discrete power transistors. Although MEMS switches have low on-state resistance and can operate up to tens of GHz with low power consumption, MEMS switches require large control voltages. PIN diode and CMOS switch-based digital TMNs exhibit low-to-moderate on-state resistance and, thus, can handle high power levels at the expense of some resistive power loss. PIN diode and CMOS switch-based digital TMNs are favorable for on-die integration, for example for Software Defined Radio (SDR) integrated circuits (ICs) and other on-chip TMNs. Digital TMNs, however, exhibit limited tuning resolution, and hence, limited accuracy with which impedance matching can be achieved. In some high power applications where accurate impedance matching is required over a very wide impedance range, such as RF plasma drivers, the use of digital TMNs can be impractical due to the largo number of digital switches needed to achieve the required fine-tuning resolution.

SUMMARY

This Summary is provided to introduce a selection of concepts in a simplified form that are further described below in the Detailed Description. This Summary is not intended to identify key or essential features or combinations of the claimed subject matter, nor is it intended to be used to limit the scope of the claimed subject matter.

It has been recognized that there is a need for TMNs having increased accuracy and/or faster impedance matching relative to existing TMNs. It has also been recognized that there is a need for TMNs having increased accuracy and/or faster impedance matching with higher tuning bandwidth over wider impedance ranges, while simultaneously allowing operation at higher power levels with low insertion losses.

One aspect of the concepts, systems and techniques described herein are directed toward a radio frequency (RF) amplifier system having an input port and an output port. The RF amplifier system includes an RF amplifier having an input port coupled to the input port of the RF amplifier system and having an output port. A phase-switched tunable impedance network is coupled between the output port of the RF amplifier and an output port of the RF amplifier system. In response to signals provided thereto, the phase-switched tunable impedance network varies an impedance thereof to modulate an impedance presented to the output port of the RF amplifier.

In another aspect of the concepts, systems and techniques described herein, a radio frequency (RF) amplifier system having an input port and an output port includes a phase-switched tunable impedance network having an input port coupled to the input port of the RF amplifier system and having an output port. An output of the phase-switched tunable impedance network is coupled to an input of an RF amplifier and an output port of the RF amplifier is coupled to the output port of the RF amplifier system. In response to signals provided thereto, the phase-switched tunable impedance network varies an impedance thereof to modulate an impedance presented to the input port of the RF amplifier.

In still another aspect of the concepts, systems and techniques described herein a radio frequency (RF) amplifier system having an input port and an output port includes a first phase-switched tunable impedance network having an input port coupled to the input port of the RF amplifier system and having an output port. An output of the first phase-switched tunable impedance network is coupled to an input of an RF amplifier and an output port of the RF amplifier is coupled to an input port of a second phase-switched tunable impedance network. An output of the second phase-switched tunable impedance network is coupled to the output port of the RF amplifier system. In response to signals provided to respective ones thereof, the first phase-switched tunable impedance network varies an impedance thereof to modulate an impedance presented to the input port of the RF amplifier and the second phase-switched tunable impedance network varies an impedance thereof to modulate an impedance presented to the output port of the RF amplifier.

In embodiments, the phase-switched tunable impedance network includes one or more phase-switched reactive ele-

ments. Each of the phase-switched reactive elements receive a respective control signal. In response to the respective control signal provided thereto, each phase-switched reactive element is provided having a corresponding desired reactance value.

In embodiments, a controller provides a respective control signal to each of the one or more phase-switched reactive elements.

In embodiments, each of the one or more phase-switched reactive elements includes one or more reactive elements and at least one switch. At least one of the one or more reactive elements is configured to be switched into and out of the phase-switched tunable impedance network by at least one switch associated therewith.

In embodiments, the at least one associated switch is operable at a switching frequency related to a frequency of an RF signal at the output port of the RF amplifier and a switching phase based upon the respective control signal.

In embodiments, the at least one switch is operable in a half-wave switching configuration to switch on and off once per cycle of the RF signal at the output port of the RF amplifier.

In embodiments, the at least one switch is operable in a full-wave switching configuration to switch on and off twice per cycle of the RF signal at the output port of the RF amplifier.

In embodiments, each of the one or more phase-switched reactive elements includes a capacitor in parallel with a switch.

In embodiments, each of the one or more phase-switched reactive elements includes an inductor in series with the combination of the capacitor in parallel with the switch.

In embodiments, the at least one switch is operable to provide at least one of: zero voltage switching and zero current switching of said at least one switch.

In embodiments, the switching frequency and the switching phase are selected to provide the phase-switched reactive element having a desired reactance value.

In embodiments, at least one of the one or more phase-switched reactive elements is a capacitive element having a capacitive value. The capacitance value of the phase-switched capacitive element at a desired frequency is related to a physical DC capacitance value of the phase-switched capacitive element and the switching phase.

In embodiments, at least one of the one or more phase-switched reactive elements is an inductive element having an inductance value. The inductance value of the phase-switched inductive element at a desired frequency is related to a physical DC inductance value of the phase-switched inductive element and the switching phase.

In embodiments, the impedance presented to the output port of the RF amplifier by the phase-switched tunable matching network is dynamically adapted to match a variable load impedance coupled to the output port of the RF amplifier system to an impedance of the RF amplifier.

In embodiments, an RF load is coupled to the output port of the RF amplifier system. The RF load is at least one of: an antenna; a transmission line; and a plasma load.

In embodiments, the RF amplifier includes a switching inverter comprising at least one switching element configured to generate RF power.

In embodiments, the controller modulates the impedance of the phase-switched tunable impedance network presented to the output port of the RF amplifier such that the RF amplifier maintains zero-voltage-switching (ZVS) of the at least one switching element of the switching inverter.

In embodiments, switching the at least one switch of each of the one or more phase switched reactive elements modulates an impedance transformation provided by the phase switched tunable impedance network.

In embodiments, the RF amplifier includes a filter coupled to the phase-switched tunable impedance network. The filter has a filter characteristic that reduces harmonic content generated by the phase-switched tunable impedance network and presented to at least one of the output port of the RF amplifier and the output port of the RF amplifier system.

In embodiments, the filter includes one or more filter components to electrically isolate DC signals between the phase-switched tunable impedance network and at least one of the RF amplifier and the output port of the RF amplifier system.

In embodiments, the phase-switched tunable impedance network includes a first shunt path phase-switched variable reactive element coupled to a first node of a series path reactive element.

In embodiments, the phase-switched tunable impedance network also includes a second shunt path phase-switched variable reactive element coupled to a second node of the series path reactive element.

In another aspect, a method of operating a radio frequency (RF) amplifier system includes providing an RF signal to an input port of an RF amplifier and amplifying the RF signal by the RF amplifier to provide an amplified RF signal at an output port of the amplifier. Varying an impedance of a phase-switched tunable impedance network coupled between the output port of the RF amplifier and an output port of the RF amplifier system modulates an impedance presented to the RF amplifier.

In embodiments, varying the impedance of the phase-switched tunable impedance network includes receiving a control signal by the phase-switched tunable impedance network. In response to the control signal, at least one reactive element is electrically connected or disconnected into or out of the phase-switched tunable impedance network at a frequency and phase related to a frequency of an RF signal at the output port of the RF amplifier to provide the phase-switched tunable impedance network having a desired reactance value.

In embodiments, the phase-switched tunable impedance network includes one or more reactive elements and at least one switch. Electrically connecting or disconnecting at least one reactive element into or out of the phase-switched tunable impedance network includes switching at least one of the one or more reactive elements into and out of the phase-switched tunable impedance network by at least one switch associated therewith and operating the at least one associated switch at a switching frequency and a switching phase based upon the respective control signal.

In embodiments, the switching frequency and the switching phase based upon the frequency of the RF signal at the output port of the RF amplifier are selected to provide the phase-switched reactive element having a desired reactance value.

In embodiments, the impedance presented to the RF amplifier by the phase-switched tunable matching network is dynamically adapted to match a variable load impedance coupled to the output port of the RF amplifier system to an impedance of the RF amplifier.

In embodiments, modulating the load impedance presented to the RF amplifier by the phase-switched tunable impedance network controls a power level of the amplified signal provided to the output port of the RF amplifier system.

In embodiments, the amplifier includes a switching inverter comprising at least one switching element configured to generate RF power. Zero-voltage-switching (ZVS) of the at least one switching element of the switching inverter is maintained by modulating the impedance of the phase-switched humble impedance network presented to the RF amplifier.

In embodiments, the at least one switch of the phase-switched tunable impedance network is switched in manner to provide at least one of: zero voltage switching and zero current switching of said at least one switch.

BRIEF DESCRIPTION OF THE DRAWING FIGURES

Other aspects, features, and advantages of the broad concepts sought to be protected herein will become more fully apparent from the following detailed description, the appended claims, and the accompanying drawings in which like reference numerals identify similar or identical elements. Reference numerals that are introduced in the specification in association with a drawing figure may be repeated in one or more subsequent figures without additional description in the specification in order to provide context for other features.

FIG. **1** is a block diagram of an illustrative tunable impedance matching network (TMN) in accordance with described embodiments;

FIG. **2** is a schematic diagram of an illustrative phase-switched variable capacitance element of the TMN of FIG. **1**;

FIG. **3** is a plot of current and voltage versus phase with respect to a control signal of the phase-switched variable capacitance element of FIG. **2**;

FIG. **4** is a schematic diagram of an illustrative phase-switched variable inductance element of the TMN of FIG. **1**;

FIG. **5** is a plot of current and voltage versus phase with respect to a control signal of the phase-switched variable inductance element of FIG. **4**;

FIG. **6** is a plot of normalized effective capacitance (or inductance) of the phase-switched elements of FIGS. **2** and **4** versus a control angle of the phase-switched element;

FIG. **7** is a plot of total harmonic distortion of the phase-switched elements of FIGS. **2** and **4** versus the control angle of the phase-switched element;

FIG. **8** is a plot of current and voltage versus phase with respect to a control signal of a full-wave switched variable capacitance element;

FIG. **9** is a plot of current and voltage versus phase with respect to a control signal of a full-wave switched variable inductance element;

FIGS. **10**A-D are schematic diagrams of illustrative switched reactance elements in accordance with described embodiments;

FIG. **11** is a schematic diagram of an illustrative phase-switched TMN employing a digitally-switched capacitance matrix;

FIG. **12** is a schematic diagram of an illustrative phase-switched TMN employing a digitally-switched inductance matrix;

FIG. **13** is a schematic diagram of an illustrative phase-switched TMN in accordance with described embodiments;

FIG. **14** is a Smith chart of a range of load impedances that can be matched by the tuning network of FIG. **13** for an illustrative operating range;

FIG. **15** is a schematic diagram of additional detail of the tuning network of FIG. **13**;

FIG. **16** is a block diagram of an illustrative topology of a phase-switched impedance modulation amplifier in accordance with described embodiments;

FIG. **17** is a block diagram of another illustrative topology of a phase-switched impedance modulation amplifier in accordance with described embodiments;

FIGS. **18**A-E are schematic diagrams of illustrative three-switch phase-switched impedance modulation amplifiers in accordance with described embodiments;

FIGS. **19** and **20** are schematic diagrams of illustrative two-switch phase-switched impedance modulation amplifiers in accordance with described embodiments;

FIG. **21** is a schematic diagram of an illustrative phase-switched impedance modulation amplifier over an illustrative operating range;

FIGS. **22** and **23** are Smith charts showing a range of load impedances that can be matched by the phase-switched impedance modulation amplifier of FIG. **21** for an illustrative operating range; and

FIG. **24** is a flow diagram of an illustrative process of operating the TMN of FIG. **1**.

DETAILED DESCRIPTION

Table 1 summarizes a list of acronyms employed throughout this specification as an aid to understanding the described embodiments:

TABLE 1

CMOS	Complementary Metal-Oxide Semiconductor	CR	Cognitive Radio
FET	Field Effect Transistor	HEMT	High-Electron-Mobility Transistor
IC	Integrated Circuit	LUT	Look Up Table
MEMS	Micro-ElectroMechanical Systems	PA	Power Amplifier
PSIM	Phase-Switched Impedance Modulation	PS-TMN	Phase-Switched Tunable impedance Matching Network
RF	Radio Frequency	SDR	Software Defined Radio
TMN	Tunable impedance Matching Network	WPT	Wireless Power Transfer
ZCS	Zero Current Switching	ZVS	Zero Voltage Switching

Described embodiments are directed toward phase-switched, tunable matching networks (PS-TMNs) and phase-switched, impedance modulation amplifiers (PSIMs). Both the phase-switched, tunable matching networks and the phase-switched, impedance modulation amplifiers include phase-switched variable network reactance elements. When configured in the context of PS-TMNs and phase-switched, impedance modulation amplifiers, such phase-switched variable network reactance elements provide rapid, high bandwidth, continuous impedance matching over a wide impedance range, while operating efficiently at high power levels without requiring high bias voltages or currents. PS-TMNs might be employed alone, or might also be employed in combination with other matching techniques such as discrete switched reactance banks.

PS-TMNs might be employed in a variety of reconfigurable and adaptive RF systems, for example, RF front ends for software-defined radio (SDR) and cognitive radio (CR) applications that operate over a wide range of frequency bands, at different bandwidths, and in accordance with a variety of communication standards. PS-TMNs might also be employed in other RF applications, such as drivers for RF plasma loads to compensate for rapid load variations, or in wireless power transfer (WPT) systems to compensate for impedance mismatches between the transmitter and receiver to maximize transferred power and/or efficiency.

The PSIMs may be operable as zero voltage switching (ZVS) radio frequency (RF) amplifiers. Such PSIM amplifiers might employ a PS-TMN to operate over a large frequency range by efficiently modulating output power over a wide frequency range and/or matching into highly variable loads (e.g., loads that are variable over a wide impedance range).

Referring to FIG. **1**, a radio frequency (RF) system **100** includes a phase-switched tunable impedance matching network (PS-TMN) **112** coupled between a source **102**, having an impedance Z_S, and a load **114**, having an impedance Z_L. In some applications, source **102**, control circuit **106** and PS-TMN **112** (and other elements of RF system **100**) are coupled to a power supply voltage (e.g., V_{DC}) and ground. Control circuit **106** is coupled to PS-TMN **112** and provides control signals to PS-TMN **112** so as to control operation of PS-TMN **112**. In response to such control signals, PS-TMN **112** provides a desired impedance transformation characteristic. It should be appreciated that control circuit **106** might be an internal component to PS-TMN **112**, or might be an external component coupled to PS-TMN **112** or some portions of control circuit **106** (or functions provided by control circuit **106** may be internal to PS-TMN **112** while other portions of control circuit **106** may be external to PS-TMN **112**.

In some embodiments, control circuit **106** controls operation of PS-TMN **112** based, at least partially, upon information received from an optional feedforward circuit **104** coupled to source **102** and/or an optional feedback circuit **110** coupled to load **114**. In some embodiments, optional feedforward circuit **104** includes adaptive predistortion circuit **107** and control circuit **106** includes look up table (LUT) **108**. For example, as will be described in greater detail below, some embodiments might employ one or more nonlinear control techniques (e.g., by control circuit **106**) to determine appropriate control signals for PS-TMN **112**, such as employing fixed or adaptable lookup tables (e.g., LUT **108**) to store predetermined control signal information, feedback (e.g., by feedback circuit **110**) and/or feedforward compensation (e.g., by feedforward circuit **104**) to adaptively adjust control signal information, or performing digital predistortion of the control signals (e.g., by predistortion circuit **107**), or other similar techniques.

PS-TMN **112** includes one or more phase-switched reactance elements **116(1)**-**116**(N). As will be described in greater detail below, phase-switched reactance elements **116(1)**-**116**(N) might be implemented using one or more capacitive elements (e.g., capacitors), one or more inductive elements (e.g., inductors), or a combination of both. Phase-switched reactance elements **116(1)**-**116**(N) can be controlled to adjust the effective impedance ($Z_{S,IN}$ and $Z_{L,IN}$) presented to the terminals of PS-TMN **112** at a desired frequency. The phase-switched reactance elements **116(1)**-**116**(N) are switched, for example by either a shunt or a series switch, and the effective impedance of the phase-switched reactance elements is controlled by adjusting the phase and/or duty-cycle of the shunt or series switch. In some embodiments, the desired frequency might be the RF frequency of operation of RF source **102** (e.g., the frequency of the signal provided from RF source **102** to PS-TMN **112**).

By modulating the effective impedance at a desired frequency of operation of RF system **100** (e.g., by adjusting the impedance of phase-switched reactive elements **116(1)**-**116**(N)), it is possible to adjust, tune, change or otherwise manipulate the impedance presented by PS-TMN **112** to source **102** and/or load **114**. For example, phase-switched reactance elements **116(1)**-**116**(N) allow PS-TMN **112** to present a desired impedance ($Z_{S,IN}$) to PS-TMN **112** from source **102** and a desired impedance ($Z_{L,IN}$) into PS-TMN **112** from load **114**.

The control signals provided to PS-TMN **112** operate to control the timing of turning on and/or of the switches of phase-switched reactance elements **116(1)**-**116**(N) with respect to the RF signal provided from source **102**. The switching provides the effective reactance values of phase-switched reactance elements **116(1)**-**116**(N) that effect the desired impedance transformation of PS-TMN **112**. Feed-forward information might include information about the effective input impedance of PS-TMN **112**, the timing of RF waveforms, specified signal levels and/or impedance levels, etc. Feedback information might include measured information about the effective load impedance and/or power reflected from the load, the timing of RF waveforms, etc.

Thus, in some embodiments, PS-TMN **112** might be employed to provide a desired impedance transformation between source **102** and load **114**. For example, PS-TMN **112** might provide an impedance match between source **102** and load **114**. Alternatively, the impedance of PS-TMN **112** might be adjusted to compensate for variations in the impedance (Z_L) of load **114** such that source **102** is coupled to a more stable impedance (e.g., $Z_{S,IN}$) provided by PS-TMN **112**.

Referring to FIG. **2**, sinusoidal current source **202**, having a current I, drives an illustrative phase-switched variable reactance **200**. The phase-switched variable reactance is here shown as including a parallel combination of a capacitor **204** and switch **206** to provide the phase-switched variable reactance as phase-switched variable capacitance **200**. Capacitor **204** has a physical capacitance C_0, and a voltage V_C. The state of switch **206** is controlled by a characteristic of signal Q. For example, switch **206** provides a low impedance signal path between its terminals (e.g., switch **206** is "on" or "closed") when signal Q has a logic high value, and switch **206** provides a high impedance signal path between its terminals (e.g., switch **206** is "off" or "open") when signal Q has a logic low value. Thus, switch **206** could be considered to switch capacitor **204** into the circuit when the switch is open (current I flows into capacitor **204**), and out of the circuit when the switch is closed (current I flows through the closed switch and bypasses capacitor **204**).

If switch **206** is always off (open), then the effective capacitance, C_{EFF}, of phase-switched variable capacitance **200** presented to source **202** is equivalent to the physical capacitance, C_0, of capacitor **204**. Alternatively, if switch **206** is always on (closed), then the low impedance path between the terminals of switch **206** effectively "shorts" capacitor **204**, and phase-switched variable capacitance **200** behaves as an infinite capacitor in the sense that the voltage across capacitor **204** remains zero irrelevant of current I. The effective capacitance, C_{EFF}, of capacitor **204** can theoretically be controlled between C_0 and infinity by controlling the conduction angle of switch **206** over an AC cycle of sinusoidal current source **202** from 0 to 2π. As used herein, a conduction angle is the angle of the sinusoidal signal at which switch **206** is turned on. The conduction angle with which the switch is turned on may be entirely determined by a switching signal Q (e.g., the switching angle) or partly determined by switching signal Q and partly by circuit waveforms such as voltage V_C and current I.

Referring to FIG. **3**, illustrative waveforms of the current I and capacitor voltage V_C (e.g., the voltage of capacitor

204) are shown with respect to the switch control signal, Q, as a function of a cycle angle θ. In particular, curve **302** shows I(θ), curve **306** shows $V_C(\theta)$ and curve **304** shows Q(θ) for a half-wave switched capacitor. As shown in FIG. **3**, every cycle of I(θ), switch **206** is turned off (opened) α radians after I(θ) transitions from negative to positive (e.g., switch **206** is on/closed until α radians into the positive half-cycle of I(θ)). Switch **206** remains off (open) until after the capacitor voltage rings down to zero. Biasing the switch into its conductive state (e.g., turning the switch on or closing the switch) after the capacitor voltage rings down to zero ensures zero-voltage-switching (ZVS) turn on of switch **206**.

If the switch includes a diode that naturally prevents the voltage from going negative, the timing of actively turning switch Q on may be relaxed, since it will naturally commutate "ON" when the switch voltage reaches zero and the active turn-on signal can be issued while the diode conducts. The capacitor C_0 across the switch provides snubbing of the turn off transition, providing zero-voltage-switching (ZVS) turn off of switch **206**.

As shown in FIG. **3**, when I(θ) is a purely sinusoidal current source, switch **206** remains off (open) until the conduction angle of the switch is reached (e.g., at 2α). Thus, for a half-wave switched capacitor, switch **206** is turned on and off once per cycle of the RF signal from source **102** (e.g., I(θ) as shown by curve **302**).

Adjusting α sets where in the cycle switch **206** turns on and off (e.g., controls the conduction angle of switch **206**) and hence controls the voltage at which the capacitor peaks. Thus, there is a relationship between the switching angle (α) and the magnitude of the fundamental component of $V_C(\theta)$ at the switching frequency. Consequently, the effective capacitance, C_{EFF}, of capacitor **204** can be represented as a function of α:

$$C_{EFF} = \frac{C_0 \cdot \pi}{\pi - \alpha + \sin(\alpha) \cdot \cos(\alpha)} \tag{1a}$$

Referring to FIG. **4**, it is also possible to implement a phase-switched variable reactance as a switched inductor network that allows continuous control of its effective inductance at the switching frequency. Such a switched inductor network is shown in FIG. **4** as phase-switched variable inductance **400** and corresponds to the topological dual of the switched capacitor network **200** shown in FIG. **2**. As shown in FIG. **4**, illustrative phase-switched variable inductance **400** includes a series combination of inductor **404** and switch **406** being driven by a sinusoidal voltage source **402** with a voltage V. Inductor **404** has a physical inductance L_0, and an inductor current I_L. The state of switch **406** is controlled by the signal Q, for example, switch **406** might be on (e.g., closed) when signal Q has a logic high value, and off (e.g., open) when signal Q has a logic low value. Thus, switch **406** could be considered to switch inductor **404** into the circuit when the switch is closed (applying voltage V to inductor **404**), and out of the circuit when the switch is open (no voltage is applied to inductor **404**).

Similarly to the switched-capacitor implementation of a phase-switched variable reactance described in regard to FIG. **2**, the effective inductance L_{EFF} of phase-switched variable inductance **400** at the switching frequency can be modulated from a base value L_0 to infinity. For example, if switch **406** is always on (closed), then the effective inductance, L_{EFF}, of phase-switched variable inductance **400** seen by source **402** is equivalent to the physical inductance, L_0, of inductor **404**. Alternatively, if switch **406** is always off (open), then inductor **404** behaves as an infinite inductor in the sense that the current through inductor **404** remains zero irrelevant of voltage V. The effective inductance, L_{EFF}, of inductor **404** can ideally be controlled between and infinity by controlling the conduction angle of switch **406** over an AC cycle of sinusoidal voltage source **402** from 0 to 2π.

Referring to FIG. **5**, illustrative waveforms of the current I and voltage V_C of capacitor **204** are shown with respect to the switch control signal, Q, as a function of a cycle angle θ. As a result of the properties of topological duality, the voltage waveform of the switched capacitor network shown in FIG. **3** is analogous to the current waveform of the switched inductor network shown in FIG. **5**, and vice versa.

In particular, curve **502** shows $I_L(\theta)$, curve **506** shows V(θ) and curve **504** shows Q(θ) for a half-wave switched inductor. As shown in FIG. **5**, every cycle of V(θ), switch **406** is turned an (closed) α radians after V(θ) transitions from negative to positive (e.g., switch **406** is off/open until α radians into the positive half-cycle of V(θ)). Switch **406** remains on (closed) until after the inductor current rings down to zero. Since the switch has an inductor in series with it, zero-current-switching (ZCS) turn on of the switch can be achieved. Turning the switch off at the time when the inductor current rings down to zero ensures zero-current-switching (ZCS) turn off of switch **406**. In duality with the capacitive circuit, utilization of diode(s) as part of switch Q can enable natural commutation (turn off) of the switch and relax detailed active timing of the turn-off moment of the switching control waveform. As shown in FIG. **5**, when V(θ) is a purely sinusoidal voltage source, switch **406** remains on (closed) until the conduction angle of the switch is reached (e.g., at 2α).

Adjusting α sets where in the cycle switch **406** turns on and off (e.g., controls the conduction angle of switch **406**) and hence controls the current at which the inductor peaks. Thus, similarly to the switched-capacitor implementation of a phase-switched variable reactance described in regard to FIG. **2**, there is a relationship between the switching angle (α) and the magnitude of the fundamental component of $I_L(\theta)$ at the switching frequency. Consequently, the effective inductance, L_{EFF}, of inductor **404** can be represented as a function of α;

$$L_{EFF} = \frac{L_0 \cdot \pi}{\pi - \alpha + \sin(\alpha) \cdot \cos(\alpha)} \tag{1b}$$

As a result of topological duality, expression (1b) for the effective inductance is the same as that of expression (1a) for the effective capacitance. Expression (1a) is consistent with the intuitive expectation for infinite effective capacitance when the switch is always in the on state (α=π) and predicts the equivalence between C_{EFF} and C_0 when the switch is permanently off (α=0). Expression (1b) is similarly consistent with the intuitive expectation for infinite effective inductance when the switch is always in the off state (α=0) and predicts the equivalence between L_{EFF} and L_0 when the switch is permanently on (α=π). Thus, in accordance with expressions (1a) and (1b), the effective capacitance C_{EFF} or the effective inductance L_{EFF} at the switching frequency can be modulated by controlling the conduction angle of the switch associated with the capacitor or inductor.

Referring to FIG. **6**, the normalized effective capacitance, C_{EFF}/C_0, or the normalized effective inductance L_{EFF}/L_0, is shown by curve **602** at the switching frequency. For the capacitive circuit this is the same thing as the normalized admittance Y_{EFF}/Y_0 while for the inductive circuit this is the same as the normalized reactance, X_{EFF}/X_0. As a result of topological duality, the normalized effective admittance Y_{EFF}/Y_0 of the phase-switched capacitor circuit of FIG. **2** is the same as the normalized reactance, X_{EFF}/X_0 of the phased-switched inductor network shown in FIG. **4**.

As shown in FIG. **6**, the normalized effective capacitance C_{EFF} (or inductance L_{EFF}) increases rapidly with α and approaches infinity as α approaches π (e.g., 180 degrees).

Referring to FIG. **7**, curve **702** shows the total harmonic distortion of the capacitor voltage (inductor current) versus α for a purely sinusoidal current (voltage) excitation source. The practical range over which C_{EFF} or L_{EFF} can be modulated depends on the amount of harmonic distortion that can be present in the network. As α increases towards π (e.g., the conduction angle of the switch increases), the ringing of the capacitor voltage V_C (e.g., curve **306**) or of the inductor current I_L (e.g., curve **502**) is limited to a shorter time period. As shown in FIG. **7**, this results in significant harmonic content of the capacitor voltage for large Y_{EFF}/Y_0 or X_{EFF}/X_0 (e.g., C_{EFF}/C_0 or L_{EFF}/L_0) ratios (e.g., the total harmonic distortion increases as α increases). The amount of harmonic distortion allowed in a given system depends on specified limits of harmonic content allowed into the source and/or load and the amount of filtering that is necessary or desired.

Note that FIG. **7** shows the harmonic distortion of the phase-switched variable reactance (e.g., the harmonic distortion of the capacitor voltage of phase-switched variable capacitance **200**, or the harmonic distortion of the inductor current of phase-switched variable inductance **400**), and not the harmonic content that is actually injected into the source and/or load of the RF system (e.g., source **102** and load **114**). In some embodiments, the phase-switched variable reactance (e.g., phase-switched variable capacitance **200** or phase-switched variable inductance **400**) includes additional filtering components (not shown in FIGS. **2** and **4**) to reduce harmonic content injected into the source and/or load (e.g., source **102** and load **114**).

As described in regard to FIGS. **3** and **5**, the phase-switched variable reactance (e.g., phase-switched variable capacitance **200** or phase-switched variable inductance **400**), are half-wave switched, where the switch is operated so that the capacitor voltage (curve **306** of FIG. **3**) and inductor current (curve **592** of FIG. **5**) are unipolar. However, other switching schemes are also possible. For example, FIGS. **8** and **9** show illustrative waveforms of the current I and voltage V with respect to the switch control signal, Q, as a function of a cycle angle θ, for the switched capacitor network shown in FIG. **3** and the switched inductor network shown in FIG. **5**, respectively.

In particular, as shown in FIG. **8**, curve **802** shows $I(\theta)$, curve **806** shows $V_C(\theta)$ and curve **804** shows $Q(\theta)$ for a full-wave switched capacitor. As shown in FIG. **9**, curve **902** shows $I_L(\theta)$, curve **906** shows $V(\theta)$ and curve **904** shows $Q(\theta)$ for a full-wave switched inductor. When phase-switched variable capacitance **200** is full-wave switched, the switch (e.g., switch **206**) is turned off twice every cycle of $I(\theta)$ (e.g., $Q(\theta)$ is zero), with the off periods being centered around the instants when the current $I(\theta)$ is zero. For a purely sinusoidal excitation current $I(\theta)$, this results in a bipolar capacitor voltage waveform $V_C(\theta)$. Capacitor voltage $V_C(\theta)$ has zero DC average value. Similarly, when phase-switched variable inductance **400** is full-wave switched, the switch (e.g., switch **406**) is turned on twice every cycle of $V(\theta)$ (e.g., $Q(\theta)$ has a logic high value), with the on periods being centered around the instants when the voltage $V(\theta)$ is zero. For a purely sinusoidal excitation voltage $V(\theta)$, this results in a bipolar inductor current waveform $I_L(\theta)$, which also has zero DC average value. Thus, for a full-wave switched capacitor (or inductor), switch **206** is tamed on and off twice per cycle of the RF signal from source **102** (e.g., $I(\theta)$ as shown by curve **802**).

As with half-wave switching (e.g., as shown in. FIGS. **3** and **5**), the effective capacitance C_{EFF} and the effective inductance L_{EFF} at the switching frequency can be modulated by controlling the switching angle, α, of the switch. The effective capacitance, C_{EFF}, of capacitor **204** can be represented as a function of α for a full-wave switched capacitor:

$$C_{EFF} = \frac{C_0 \cdot \pi}{2 \cdot [\pi - \alpha + \sin(\alpha) \cdot \cos(\alpha)]} \tag{2a}$$

Similarly, the effective inductance, L_{EFF}, of inductor **404** can be represented as a function of α:

$$L_{EFF} = \frac{L_0 \cdot \pi}{2 \cdot [\pi - \alpha + \sin(\alpha) \cdot \cos(\alpha)]}. \tag{2b}$$

Thus, the effective capacitance/inductance that can be achieved for a given switching angle, α, with full-wave switched networks (e.g., relationships (2a) and (2b)) is half the effective capacitance/inductance that can be achieved with half-waved switched networks (e.g., relationships (1a) and (1b)). However, full-wave switched networks inherently result in reduced harmonic content of the capacitor voltage and inductor current compared to half-wave switched networks for the same switching angle, α (i.e the switching angle which controls the total switch conduction angle). On the other hand, implementing full-wave switching requires the switch has to operate at twice the operating frequency (e.g., to switch twice per cycle). Further, for capacitive modulation, bidirectional blocking switches are required, which can complicate switch implementation with typical semiconductor switches.

Relationships (1) and (2) above show that the effective capacitance and inductance for the switched networks shown in FIGS. **2** and **4** can be based upon the switching angle, α, for a purely sinusoidal excitation signals. For excitation signals that are not purely sinusoidal, the effective reactance can be controlled by appropriately selecting the timing or switching angle, α, at which the switch turns off (or on) although relationships (1) and (2) cannot calculate an exact value of α. Together with the circuit waveforms that determine zero-voltage (or zero current) points (for switch turn on (or off), switching angle α determines the total conduction angle of the switch during the cycle. For excitation signals that are not purely sinusoidal, an adaptable look-up table (e.g., LUT **108**), feedback circuit **110** or feedforward circuit **104** (including optional digital predistortion circuit **107**) might be employed to determine the required value of α for a given desired effective reactance.

Phase-switched variable capacitance **200** and phase-switched variable inductance **400** can be employed as building blocks for implementing phase-switched variable reactances and other adjustable circuits such as TMNs. Particularly, some applications could benefit substantially from variable reactances whose value can be controlled over a range spanning both capacitive and inductive reactances, and/or by modulating the effective reactance over a more limited range. Augmenting phase-switched variable capacitance **200** and/or phase-switched variable inductance **400** with additional reactive components can provide a wider range of variable reactances.

FIGS. **10**A-**10**D show illustrative embodiments of phase-switched reactance circuits that include both capacitive and inductive elements, thereby expanding a range over which the impedance of the phase-switched reactance circuit can be tuned as compared to the single-element circuits shown in FIGS. **2** and **4**.

For example, FIG. **10**A shows phase-switched reactance circuit **1002** that includes inductor **1012** in series with phase-switched capacitor **1013**. Phase-switched capacitor **1013** includes switch **1016** in parallel with capacitor **1014**, similarly as described in regard to FIG. **2**. FIG. **10**B shows phase-switched reactance circuit **1004** that includes inductor **1024** in series with capacitor **1022**, with the series combination of inductor **1024** and capacitor **1022** arranged in parallel with phase-switched capacitor **1025**. Capacitor **1022** is not phase-switched and, thus, is shown as C_{DC}. Phase-switched capacitor **1025** includes switch **1028** in parallel with capacitor **1026**, similarly as described in regard to FIG. **2**. FIG. **10**C shows phase-switched reactance circuit **1006** that includes capacitor **1032** in parallel with phase-switched inductor **1033**. Phase-switched inductor **1033** includes switch **1036** in series with inductor **1034**, similarly as described in regard to FIG. **4**. FIG. **10**D shows phase-switched reactance circuit **1008** that includes inductor **1042** in parallel with capacitor **1044**, with the parallel combination of inductor **1042** and capacitor **1044** arranged in series with phase-switched capacitor **1045**. Inductor **1042** is not phase-switched and, thus, is shown as L_{DC}. Phase-switched inductor **1045** includes switch **1048** in series with inductor **1046**, similarly as described in regard to FIG. **4**.

As would be understood by one of skill in the art, circuit variants other than the ones illustrated in FIGS. **10**A-**10**D are also possible. For example, placing a capacitor in series with a phase-switched capacitor provides a net effective impedance having a maximum capacitance equal to the series combination of the capacitor and the physical capacitance of the phase-switched capacitor, and a minimum capacitance equal to the series combination of the capacitor and the phase-switched capacitance value.

As described in regard to FIGS. **6** and **7**, a tradeoff exists for phase-switched variable capacitance **200** and phase-switched variable inductance **400** between their variable reactance range and the amount of harmonic content injected into the rest of the system. In other words, the range over which the effective reactance can be controlled is limited by the amount of harmonic content that can be tolerated within the system (e.g., by source **102** and/or load **114**). Some embodiments might employ additional or external filtering components to reduce harmonic content injected to source **102** and/or load **114**. However, in some embodiments, it might not be possible to employ additional filtering components.

Referring to FIGS. **11** and **12**, in cases where additional filtering components are not employed, the harmonic content can be reduced by combining phase-switched variable capacitance **200** and phase-switched variable inductance **400** with one or more digitally controlled capacitor or inductor matrices that are not phase-switched. Such hybrid switched networks include an RF switch operated at the RF frequency of operation and with controlled phase and duty cycle with respect to the RF waveform. The hybrid switched network also includes digital switches associated with one or more capacitors or inductors in the switched matrix. The digital switches are typically operated at a much lower frequency than the RF frequency, but could be operated up to the RF frequency (e.g., on a cycle-by-cycle basis) determined by the control bandwidth of the effective reactance C_{EFF} or L_{EFF}.

Referring to FIG. **11**, hybrid switched network **1100** includes a phase-switched reactance (e.g., capacitor C_0 **1116** and parallel switch **1118**) and digitally controlled capacitor network **1102**. Although shown as a phase-switched variable capacitance (e.g., capacitor C_0 **1116** and parallel switch **1118**) coupled in parallel with digitally controlled capacitor network **1102** and load **114**, in other embodiments, the phase-switched reactance might be implemented as a phase-switched variable inductance (e.g., such as shown in FIG. **4**) coupled in series with digitally controlled capacitor network **1102** and load **114**, or as one of the phase-switched reactance circuits shown in FIGS. **10**A-D, or other equivalent circuits.

Digitally controlled capacitor network **1102** includes a plurality of capacitors and associated switches, shown as capacitors **1104**, **1108**, and **1112**, and switches **1106**, **1110**, and **1114**. In some embodiments, each of capacitors **1104**, **1108**, and **1112** have a unique capacitance value, allowing the capacitance value of digitally controlled capacitor network **1102** to be varied across a large capacitance range. For example, as shown in FIG. **11**, capacitors **1104**, **1108**, and **1112** might increase from the phase-switched capacitor base value (e.g., C_0) in increments of C_0 until reaching a maximum capacitance value (e.g., $(2 \cdot 2^N - 1) \cdot C_0$), where N is the number of capacitors in digitally controlled capacitor network **1102**).

Switches **1106**, **1110**, and **1114** are coupled in series with corresponding ones of capacitors **1104**, **1108**, and **1112** and are operable to adjust the capacitance of digitally controlled capacitor network **1102** by connecting (or disconnecting) the respective capacitors. Switches **1106**, **1110**, and **1114** might operate based upon one or more control signals from control circuit **106**. As described, switches **1106**, **1110**, and **1114** generally operate at a frequency less than the RF frequency to adjust the capacitance value of digitally controlled capacitor network **1102**.

Referring to FIG. **12**, hybrid switched network **1200** includes a phase-switched reactance (e.g., inductor L_0 **1216** and series switch **1218**) and digitally controlled inductor network **1202**. Although shown as a phase-switched variable inductance (e.g., inductor L_0 **1216** and series switch **1218**) coupled in series with digitally controlled inductor network **1202** and in parallel with load **114**, in other embodiments, the phase-switched reactance might be implemented as a phase-switched variable capacitance (e.g., such as shown in FIG. **2**), or as one of the phase-switched reactance circuits shown in FIGS. **10**A-D, or other equivalent circuits.

Digitally controlled inductor network **1202** includes a plurality of inductors and associated switches, shown as inductors **1206**, **1210**, and **1214**, and switches **1204**, **1200**, and **1212**. In some embodiments, each of inductors **1206**, **1210**, and **1214** have a unique inductance value, allowing the inductance value of digitally controlled inductor network **1202** to be varied across a large inductance range. For example, as shown in FIG. **12**, inductors **1286**, **1210**, and **1214** and **1218** might increase from the phase-switched inductor base value (e.g., L_0) by increments of L_0 until reaching a maximum inductance value.

Switches **1204**, **1208**, and **1212** are coupled in parallel with corresponding ones of inductors **1206**, **1210**, and **1214** and are operable to adjust the inductance of digitally controlled inductor network **1202** by connecting (or shorting, e.g., providing a low-impedance path to bypass the inductor) the respective inductors. Switches **1204**, **1208**, and **1212** might operate based upon one or more control signals from control circuit **106**. As described, switches **1204**, **1208**, and **1212** generally operate at a frequency less than the RF frequency to adjust the capacitance value of digitally controlled inductor network **1202**.

Digitally controlled capacitor network **1102** and digitally controlled inductor network **1202** expand the range over which the reactance of the phase-switched reactance (e.g., capacitor C_0 **1116** and parallel switch **1118**, or inductor L_0 **1216** and series switch **1218**) can be continuously varied without introducing excessive harmonic content to source **102** and/or load **114**. For example, the embodiments shown in FIGS. **11** and **12** employ digitally controlled capacitor network **1102** (or digitally controlled inductor network **1202**) to control the base value C_0 (or L_0) of the switched networks **1100** (or **1200**). The switch of the phase-switched reactance (e.g., switch **1118** or switch **1218**) can be operated to step-up the base capacitance C_0 (or inductance L_0) by a factor determined by relationships (1) and (2) described above.

For example, the effective capacitance C_{EFF} at the switching frequency of hybrid switched capacitor network **1100** can be controlled between a lower capacitance vale C_0 and an upper capacitance value by half-wave switching the RF switch with the switching angle, α, varying, from 0 to approximately $\pi/2$ as shown in FIG. **3**. As shown in FIG. **7**, RF switch operation with a switching angle, α, less than $\pi/2$ (90 degrees) corresponds to a peak harmonic distortion of less than approximately 35%. Thus, the hybrid switched networks (e.g., **1100** and **1200**) allow continuous control of the effective reactance at the switching frequency over a wide capacitive (or inductive) range with minimum harmonic distortion and without the need for adjustable bias voltages or currents.

In various embodiments, the RF switch of TMN **112** (e.g., switch **206** or switch **406**) can be implemented as one of or a combination of various types of switching elements, for example based upon the RF frequency or other operating parameters of RF system **100**. For example, lateral or vertical FETs, HEMTs, thyristors, diodes, or other similar circuit elements might be employed.

Phase-switched variable capacitance **200** and phase-switched variable inductance **400** can be employed as circuit elements within more complex phase-switched tunable matching networks (PS-TMNs), for example a Pi-network topology PS-TMN (Pi-TMN), although other network topologies are possible, such as L-networks, T-networks, or other similar networks. FIG. **13** shows a schematic of illustrative RF system **1300** including an RF source **1301** coupled to Pi-TMN **1302**, which is coupled to an RF load **1303**. Pi-TMN **1302** includes two variable shunt capacitive susceptances B_1 **1310** and B_2 **1314**. In illustrative embodiments, RF source **1301** is commonly a power amplifier or the output of another RF system. As shown in FIG. **13**, RF source **1301** can be represented by its Norton equivalent circuit as including current source **1304** in parallel with source resistance R_S **1306** and source susceptance B_S **1308**. Similarly, RF load **1303** can be represented as including load resistance R_L **1318** in parallel with load susceptance B_L **1316**. The source and load impedances, Z_S and Z_L, respectively, can be expressed as:

$$Z_S=(R_S^{-1}+jB_S)^{-1} \tag{3}$$

$$Z_L=(R_L^{-1}+jB_L)^{-1} \tag{4}.$$

Thus, it can be shown that the susceptances B_1 and B_2 required to match the load impedance Z_L to the source impedance Z_S are given by:

$$B_1 = \frac{R_S \pm \sqrt{R_L R_S - X^2}}{X} - B_S \tag{5}$$

$$B_2 = \frac{R_S}{R_L}\left(\frac{R_S \pm \sqrt{R_L R_S - X^2}}{X}\right) - B_L. \tag{6}$$

Thus, Pi-TMN **1302** can be employed to match load impedance Z_L to source impedance Z_S by adjusting the values of variable shunt capacitive susceptances B_1 **1310** and B_2 **1314**.

As shown in FIG. **13**, embodiments of Pi-TMN **1302** include two variable shunt capacitive susceptances B_1 **1310** and B_2 **1314**, and a fixed inductive reactance X **1312**, although numerous other implementations of a Pi-TMN are possible, such as employing variable shunt inductive susceptances and a fixed capacitive reactance, implementing all three reactive branches as variable components, etc. It should, of course, be appreciated that is also possible to realize an L-section TMN having one variable shunt-path element and one variable series-path element. Other types of networks, might also be employed. As described in greater detail below, ground-referenced variable capacitors are highly suitable for realization with phase-switched variable reactance networks at RF frequencies.

Referring to FIG. **14**, an illustrative range of load impedances that can be matched by Pi-TMN **1302** is shown as shaded region **1402** in Smith chart plot **1400** (normalized to R_S). For example, the impedance values represented by shaded region **1402** might be achieved by an illustrative Pi-TMN having $X=R_S$ and susceptance B_1 variable over a range of $1/R_S$ to $4/R_S$, and susceptance B_2 variable over a range of $1/R_S$ to $2/R_S$. As shown in FIG. **14**, Pi-TMN **1302** is able to match the impedance of RF source **1301** to a load impedance that varies over approximately a 10:1 resistance range and a 5:1 reactance range (both capacitively and inductively). To do so, Pi-TMN **1302** modulates B_1 over as 1:4 range and B_2 over a 1:2 range, which can be achieved employing a phase-switched variable reactance network such as shown in FIGS. **2** and **4**.

FIG. **15** shows an illustrative embodiment of phase-switched Pi-TMN circuit **1502** to achieve the matching range shown in FIG. **14** for a source impedance (e.g., R_S **1506**) of 50Ω. The inductive reactance X is chosen to be equivalent in value to the Norton-equivalent source resistance R_S (e.g., 50Ω). As shown in FIG. **15**, the variable capacitive susceptances B_1 and B_2 are implemented as half-wave phase-switched capacitors (e.g., phase-switched capacitor **200** of FIG. **2**). Variable capacitive susceptance B_1 includes phase-switched capacitor C_{P2} **1514** and FET switch **1512**, which is controlled by switching control signal **q2**, which has a switching angle, α_2. Variable capacitive susceptance B_2 includes phase-switched capacitor C_{P1} **1520** and FET switch **1522**, which is controlled by switching control signal **q1**, which has a switching angle, α_1.

In an illustrative embodiment, phase-switched Pi-TMN circuit **1502** operates at 27.12 MHz and is capable of matching a 50Ω source impedance to a load impedance that varies over approximately a 10:1 resistance range and a 5:1 reactance range (both capacitively and inductively), by properly adjusting the switching angles (α_1 and α_2) of the switches and the phase shift between them (e.g., by adjusting switching control signals q**1** and q**2**).

Implementing variable capacitive susceptances B_1 and B_2 as half-wave FET-switched capacitor networks provides zero-voltage-switched (ZVS) operation of the switches, and allows each variable reactance to be implemented with a single, ground-referenced switch (e.g., FET **1512** for variable capacitive susceptance B_1 and FET **1522** for variable capacitive susceptance B_2). ZVS operation is desired in switched systems as it reduces switching power loss and improves the overall system efficiency. Furthermore, the output (drain-to-source) capacitance of FETs **1512** and **1522** are in parallel with phase-switched capacitors C_{P1} and C_{P2} and, thus can be added to the shunt capacitances and utilized as part of the TMN.

In illustrative Pi-TMN circuit **1502**, inductive reactance X **1312** shown in FIG. **13** is implemented as a series-resonant circuit including inductor L_{S2} **1516** and capacitor C_{S2} **1518** disposed in series between variable susceptances B_1 and B_2, which are disposed as shunt elements (e.g., coupled to ground). Inductor L_{S2} **1516** and capacitor C_{S2} **1518** are selected to have an inductive impedance approximately equal to the source impedance (e.g., 50Ω) at the desired frequency.

In the embodiment shown in FIG. **15**, two additional series-resonant circuits are included, one as an input filter and one as an output filter of Pi-TMN circuit **1502** to limit the amount of harmonic content injected into the source and the load as a result of the switching. For example, capacitor C_{S1} **1508** and inductor L_{S1} **1510** act as a series-resonant input filter between source **1504** and Pi-TMN circuit **1502**. Similarly, inductor L_{S3} **1524** and capacitor C_{S3} **1526** act as a series-resonant output filter between load **1528** and Pi-TMN circuit **1502**.

The quality factor, Q, of the series-resonant circuit of L_{S2} **1516** and C_{S2} **1518** controls the interaction between phase-switched capacitor C_{P1} **1520** and phase-switched capacitor C_{P2} **1514**. For example, increasing the quality factor Q (e.g., by increasing the values of L_{S2} **1516** and C_{S2} **1518**) reduces the interaction between phase-switched capacitor C_{P1} **1520** and phase-switched capacitor C_{P2} **1514**, although increasing the quality factor Q also reduces the effective bandwidth of the network.

For example, for phase-switched Pi-TMN circuit **1502** to achieve the matching range shown in FIG. **14** for a source impedance (e.g., R_S **1506**) of 50Ω at an illustrative desired frequency in the range of about 27 MHz, phase-switched capacitor C_{P1} **1520** might have a physical value, C_0, of 130 pF, and phase-switched capacitor C_{P2} **1514** might have a physical value, C_0, of 100 pF. To achieve the desired quality factor Q by the series-resonant circuit between phase-switched capacitor C_{P1} **1520** and phase-switched capacitor C_{P2} **1514**, capacitor C_{S2} **1518** might have a value of 0.01 μF, and inductor L_{S2} **1516** might have a value of 297 nH. To achieve the desired input and output filtering by the series-resonant circuits, capacitors C_{S1} **1508** and C_{S3} **1526** might have a value of 23.4 pF, and inductors L_{S1} **1510** and L_{S3} **1524** might have a value of 1.47 μH. Further, FETs **1512** and **1522** might have an on-state resistance of 10 mΩ and the body diode of each FET might have a forward voltage of 0.4V and an on-state resistance of 10 mΩ.

Switching of FETs **1512** and **1522** is synchronized to their drain current based upon the switching angle α, which is based upon the desired effective capacitance of capacitors C_{P1} and C_{P2}. As described above for half-wave phase-switched capacitors, FETs **1512** and **1522** are turned off after their drain current crosses from negative to positive, and they turned on again once their respective drain voltages ring down to zero. The appropriate value of α for each of FETs **1512** and **1522** can be calculated by determining the required B_1 and B_2 susceptances for a desired load impedance Z_L as given by relationships (5) and (6). Once each capacitive susceptance B_1 and B_2 is known, that value can be plugged in as C_{EFF} (C_0 is a known value as the physical capacitance of the capacitor) in relationship (1a) (for a half-wave phase-switched capacitor) or relationship (2a) (for a full-wave phase-switched capacitor) to determine values of α that correspond to the desired susceptance values.

As described, for phase-switched networks having non-purely sinusoidal current excitation, relationships (1) and (2) might not result in an exact value of α to achieve the desired susceptance. Further, nonlinearity of the drain-to-source switch capacitances and the mutual interaction of the two switched networks (e.g., capacitive susceptances B_1 and B_2) might also result in inaccurate calculation of α. Thus, some embodiments employ non-linear control techniques (e.g., by control circuit **106**) to determine the appropriate values of α, such as fixed or adaptable look-up tables (e.g., LUT **108**), feedback (e.g., by feedback circuit **110**), feedforward compensation (e.g., by feedforward circuit **104**), digital predistortion of the switching angles (e.g., by predistortion circuit **107**), or other similar techniques.

To set the correct value of switching control parameter α for each of FETs **1512** and **1522** for Pi-TMN circuit **1502** to achieve a given impedance, LUT **108** might store predetermined switching angles (e.g., α_1 and α_2) corresponding to various load impedances. For example, table 2 shows an illustrative list of possible load impedances that Pi-TMN circuit **1502** can match to a 50Ω source and the corresponding values of switching angles α_1 and α_2 for the switch control signals q**1** and q**2**:

TABLE 2

Load Impedance Z_L (Ω)	α_1 (degrees)	α_2 (degrees)
48.8 + 10.90j	0.0	0.0
103 + 8.12j	78.1	95.7
165 − 0.923j	87.9	91.8
282 + 3.20j	97.6	85.9
524 − 19.30j	107.0	79.1
1000 + 15.90j	117.0	72.2

Table 2 shows that it is possible for Pi-TMN circuit **1502** to match a 50Ω source impedance to a load impedance that varies resistively over at least a factor of 10:1. Based upon the switching angles (α_1 and α_2) listed table 2 and the plot of effective reactance (e.g., C_{EFF}/C_0 or L_{EFF}/L_0) versus or shown in FIG. **6**, it can be shown a 2:1 modulation of the effective capacitances can achieve impedance matching for a load impedance varying resistively over a 10:1 range.

Other types of systems can also employ the phase-switched networks described herein. For example, a wide range of systems can benefit from an RF power amplifiers (PA) that deliver power at a particular frequency or over a particular band of frequencies. Such PAs might beneficially control output power over a wide range and maintain high efficiency across its operating range. Conventional linear amplifiers (e.g., class A, B, AB, etc.) offer the benefits of wide-range dynamic output power control and high fidelity amplification, but have limited peak efficiency that degrades rapidly with power back-off. On the other hand, switching

PAs (e.g., inverters such as class D, E, F, Φ, etc.), offer high peak efficiency, but only generate constant envelope signals (at a constant supply voltage) while remaining in switched mode.

One technique for output power control in a switching PA is through load modulation, where the load of the PA is modulated by an external network. In described embodiments, the load of the PA is modulated by a phase-switched tunable matching network (TMN) (e.g., a network including one or more phase-switched variable capacitances **200** or phase-switched variable inductances **400**, such as Pi-TMN circuit **1502**). For example, an impedance transformation of a phase switched TMN might control the output power of a PA.

Referring to FIG. **16**, such a phase-switched impedance modulation (PSIM) amplifier is shown as PSIM amplifier **1600**. PSIM amplifier **1600** includes RF power amplifier (or inverter) **1602** that generates RF power at a particular frequency, or over a particular range of frequencies. RF PA **1602** is coupled to a power supply (e.g., voltage V_{DC} and ground) and phase-switched TMN **1604**. Phase-switched TMN **1604** is coupled to RF load **1606**, which has a load impedance Z_L. Phase-switched TMN **1604** is coupled to controller **1608**, which controls operation of the TMN, for example by providing control signals to switches of the TMN based upon the switching angles (e.g., α) to achieve a desired impedance. Although not shown in FIG. **16**, in some embodiments, controller **1608** is coupled to RF PA **1602** and also controls operation of the PA. Phase-switched TMN **1604** adaptively controls transforming the load impedance Z_L to an impedance presented to PA **1602**. For example, phase-switched TMN **1604** may control the output power of PA **1602** by modulating the load presented to PA **1602** (e.g., Z_{TMN}) and/or to compensate for frequency and/or load impedance variations to provide high efficiency and desired power to the load.

In various embodiments, PA **1602** is (1) a switching inverter, (2) an amplitude-modulated linear PA, or (3) a combination of these (e.g., depending on desired output). For example, FIG. **17** shows a block diagram of illustrative PSIM amplifier, **1700**, that includes switching PA **1702** (e.g., a class E, F or Φ PA, etc.) that includes a single switch (e.g., FET **1706**). In other embodiments, other types of PAs might be employed, such as linear PAs (e.g., class A, B, AB or C) or other switching PAs that use more than one switch to convert DC power to RF power (e.g., class D, inverse-D, etc.).

As described, modulating the effective loading impedance Z_{TMN} seen by the PA looking into the phase-switched TMN (e.g., TMN **1604** or **1710**) controls the output power over the operating power range of the PSIM amplifier (e.g., amplifiers **1602** and **1702**). Additionally, the operating power range of the PSIM amplifier can be further extended by also employing amplitude modulation of the PA drive signal for large output power back off.

Some embodiments might also employ other power modulation techniques such as discrete or continuous drain modulation of the power amplifier. Drain modulation of the PA modulates (e.g., switches) a bias voltage applied to a bias terminal of the PA. For example, one drain modulation technique might switching the bias voltage among multiple discrete voltage levels or continuously adjusting the bias voltage across a voltage range.

In addition to performing impedance modulation and output power control of the RF PA, a phase-switched TMN (e.g., TMN **1604** or **1710**) can also compensate for variation in the load impedance Z_L. For example, the phase-switched TMN can be continuously tuned to match a variable load impedance to a desired RF inverter loading impedance, Z_{TMN}, for a given output power level, by employing the phase-switched TMN to compensate for variations in the amplifier's loading network impedance as the operating frequency changes and, thus, maintain ZVS operation. Thus, a PSIM amplifier (e.g., PSIM amplifiers **1600** and **1700**) dynamically controls the output power it delivers to a widely varying load impedance, such as an RF plasma load, across a large frequency range.

Therefore, a PSIM amplifier (e.g., PSIM amplifiers **1600** and **1700**) allows (1) efficient dynamic control of output power over a wide power range; (2) the ability to impedance match and deliver power into a wide-ranging load, and (3) fully zero-voltage-switching (ZVS) operation across a frequency range for frequency-agile operation.

Although the block diagrams of PSIM amplifiers **1600** and **1700** shown in FIGS. **16** and **17** show PSIM amplifiers as a cascade combination of an RF PA (e.g., RF PAs **1602** and **1702**) with a phase-switched TMN (e.g., phase-switched TMNs **1604** and **1710**), other embodiments integrate the PS-TMN into the design of the RF PA. As a result, such integrated PSIM amplifiers can be viewed as an RF amplifier including two or more switches, where a first switch (or group of switches) is principally responsible for generating RF power from DC input power, and a second switch (or group of switches) is principally responsible for modulating the effective impedance presented by a load network to the RF amplifier. In most embodiments, the second switch (or group of switches) will not convert DC power to RF power (e.g., the second switch provides zero power conversion from DC to RF), although in some embodiments, the second switch may convert some power from DC to RF or RF to DC.

In most embodiments a PSIM amplifier can be a zero-voltage switching (ZVS) amplifier with the switching transistors operating substantially in switched-mode and turning on and off under zero-voltage switching, enabling high efficiency to be achieved. In other implementations, a PSIM amplifier might provide switched-mode operation (e.g., saturated operation) over some of its operating range (e.g., while delivering high output power) and utilize linear-mode operation over other portions of its range.

For example, FIG. **18**A shows a schematic of an illustrative topology for PSIM amplifier **1800**A. As shown, PSIM amplifier **1800**A is coupled to DC source **1802** coupled in series with inductor L_F, which is in turn coupled to the parallel combination of transistor **1804** and capacitor C_F. Inductor L_F, capacitor C_F, and FET **1804** generally operate to generate RF output power to the rest of the network from the DC source. Branch reactance X_1 is coupled between capacitor C_F and node N_2, which is coupled to a Pi-TMN including reactance X_2 coupled between a first phase-switched reactance (e.g., FET **1806**, branch reactance X_{S2}, and phase-switched variable reactance X_{P2}) and a second phase-switched reactance (e.g., FET **1808**, branch reactance X_{S3}, and phase-switched variable reactance X_{P3}). Branch reactance X_1 is coupled between the Pi-TMN at node N_1 and the load impedance Z_L. The branch reactances X_1, X_2, X_3, X_{S2}, X_{S3}, and the phase-switched variable reactances X_{P2} and X_{P3} can be implemented as various different reactive networks depending on the required functionality of the design.

FIG. **18**B shows an illustrative design **1800**B of the PSIM amplifier topology shown in FIG. **18**A. As shown in FIG. **18**B, the phase-switched variable reactances (comprising FET switches **1806** and **1808** and phase-switched capacitors

C_{P2} and C_{P3}) are implemented with half-wave phase-switched capacitor network such as described in regard to FIGS. **2** and **3**. As shown in FIG. **18**B, the three switches **1814**, **1816** and **1818** are mutually isolated at DC (e.g., by capacitors C_{S1}, C_{S2} and C_{S3}, respectively. FET switch **1814** is responsible for generating all the RF power, while FET switches **1816** and **1818** are responsible for transforming and modulating the impedance presented by load Z_L to the DC-to-RF portion of the circuit (e.g., at the output port of switch **1814** at node N_2).

FIG. **18**C shows an illustrative design **1800**C of the PSIM amplifier topology shown in FIG. **18**A. Network **1800**C is similar to network **1800**B, although in network **1800**C, the phase-switched capacitor networks (e.g., FET **1826** and capacitor C_{P2} and FET **1828** and capacitor C_{P3}) are connected in series with capacitors C_{P4} and C_{P5}, respectively. This decreases the sensitivity of the PSIM amplifier to variations in the effective reactance of the switched capacitor networks.

FIG. **18**D shows an illustrative design **1800**D of the PSIM amplifier topology show in FIG. **18**A where FET switches **1834** and **1836** are DC coupled (e.g., via inductor L_{S1}), and thus, potentially, one or both of FET switches **1834** and **1836** can be used to convert DC power into RF power or vice-versa. FET switch **1838**, on the other hand, is DC-isolated (e.g., by capacitors C_{S2} and C_{S3}) and, thus, is used only for impedance matching to the load impedance Z_L.

FIG. **18**E shows an illustrative design **1800**E of the PSIM amplifier topology shown in FIG. **18**A where all three FET switches **1844**, **1846** and **1848** are DC coupled (e.g., via inductor L_{S2}), while only the load is DC-isolated (e.g., by capacitor C_{S3}). Thus, in such an embodiment, all three FET switches **1844**, **1846** and **1848** can potentially be used to convert between DC power and RF power and/or be responsible for impedance matching of the network to the load, though it is not necessary that all three provide each function.

As shown in FIG. **18**E, the switched capacitor network of capacitor C_F and FET switch **1844** is in parallel with the phase-switched network of capacitor C_{P2}, inductor L_2 and FET switch **1846**. As a result, some embodiments could combine these two networks into a single switched reactive network having an input current that matches the sum of the input currents of the two switched reactive networks associated with FETs **1844** and **1846**. Thus, in some embodiments, the three-switch PSIM shown in FIG. **18**E can be implemented as a two-switch PSIM, such as shown in FIGS. **19** and **20**.

Referring to FIG. **19**, a schematic of an illustrative topology for two-switch PSIM **1900** is shown. Two-switch PSIM **1900** is coupled to RF source **1902** coupled in series with inductor L_F, which is in turn coupled to the parallel combination of FET **1904** and capacitor C_F. Branch reactance X_1 is coupled between capacitor C_F and a phase-switched reactance network including reactance X_{S2} coupled in series with the parallel combination of phase-switched reactance X_{P2} and FET **1906**. Branch reactance X_2 is coupled between the phase-switched reactance network and the load impedance Z_L. The branch reactances X_1, X_2 and X_{S2}, and the phase-switched variable reactance X_{P2} can be implemented as various different reactive networks depending on the required functionality of the design. Either one of switch FETs **1904** and **1906**, or both of switches **1904** and **1906**, can be used to convert between DC power and RF power.

Referring to FIG. **20**, an illustrative implementation of two-switch PSIM **1900** is shown having branch reactance X_1 implemented as inductor L_{S1} and capacitor C_{S1}. Capacitor CS**1** provides DC isolation between FET switches **2004** and **2006**. Thus, FET switch **2004** generates RF power and FET switch **2006** modulates the impedance presented to the source.

FIG. **21** shows an illustrative implementation of a three-switch PSIM amplifier **2100**. PSIM amplifier **2100** operates over a 20.86 MHz to 27.12 MHz frequency range (a factor of 1.3 in frequency). Further, PSIM amplifier **2100** provides the ability for 10:1 dynamic control of the output power delivered to the load having an impedance, Z_L, of 50Ω with a ±10% impedance variation (resistive and reactive).

PSIM amplifier **2100** includes RF PA (inverter) **2102**, Pi-TMN **2104**, branch filter **2106**, and load impedance Z_L. RF PA **2102** includes FET switch **2108**, inductor L_F and an output network formed by capacitors C_F and C_{S1} and inductor L_{S1}. In the embodiment shown in FIG. **21**, RF PA **2102** is a modified class E inverter with FET switch **2108** converting between DC power and RF power. Pi-TMN **2104** includes a first phase-switched capacitor (e.g., C_{P2} and FET **2110**) and a second phase-switched capacitor (e.g., C_{P1} and FET **2112**). Branch filter **2106** includes inductor L_{S3} and capacitor C_{S3} coupled between Pi-TMN **2104** and load Z_L.

RF PA **2102** maintains zero-voltage-switching (ZVS) and high efficiency at different output power levels when Pi-TMN **2104** maintains the inverter load impedance Z_{TMN} as an approximately resistive load at the frequency of operation of RF PA **2102**. RF PA **2102** generates peak RF power when Z_{TMN} is 50Ω (e.g., matches load impedance Z_L). Dynamic control of power back off of RF PA **2102** can be achieved by Pi-TMN **2104** modulating Z_{TMN}.

For operation over a 20.86 MHz to 27.12 MHz frequency range, the illustrative embodiment of PSIM amplifier **2100** shown in FIG. **21** employs inductor L_F having a value of 113 nH, capacitor C_F having a value of 180 pF, capacitor C_{S1} having a value of 15.2 pF, inductor L_{S1} having a value of 3.81 μH, phase-switched capacitor C_{P2} having a physical value, C_0, of 152 pF, inductor L_{S2} having a value of 381 nH, capacitor C_{S2} having a value of 0.01 μF, phase-switched capacitor C_{P1} having a physical value, C_0, of 152 pF, inductor LS**3** having a value of 3.81 μH, and capacitor C_{S3} having a value of 15.2 pF. In some embodiments, Pi-TMN **2104** employs half-wave switched capacitor networks (e.g., capacitor C_{P2} and FET **2110** and capacitor C_{P1} and FET **2112**).

The series reactive network branch formed by capacitor C_{S2} and inductor L_{S2} has a 50Ω inductive impedance at a frequency of 20.86 MHz and also DC isolates the two switched networks (e.g., capacitor C_{P2} and FET **2110** and capacitor C_{P1} and FET **2112**). The impedance of capacitor C_{S2} and inductor L_{S2} sets the resistive range over which Z_{TMN} of Pi-TMN **2104** can be modulated. The series resonant network formed by capacitor C_{S3} and inductor L_{S3} provides additional filtering of the load current I_L and prevents DC currents and high-frequency harmonic content being coupled to the load Z_L. Pi-TMN **2104** can modulate the impedance, Z_{TMN}, presented to RF PA **2102** by appropriately driving FET switches **2110** and **2112**, for example by adjusting the conduction angles of the FETs. By modulating the impedance Z_{TMN} presented to RF PA **2102**, Pi-TMN **2014** can control the output power that is delivered from RF PA **2102** to load Z_L.

FIG. **22** shows an illustrative impedance range (e.g., shaded region **2202**) over which Z_{TMN} of Pi-TMN **2104** can be adjusted at 20.86 MHz. FIG. **23** shows an illustrative impedance range (e.g., shaded region **2302**) over which Z_{TMN} of Pi-TMN **2104** can be adjusted at 27.12 MHz. Smith charts **2200** and **2300** are normalized to 50Ω. Shaded regions **2202** and **2302** illustrate that Pi-TMN **2104** can match load impedance Z_L over a 10:1 range by varying phase-switched capacitor C_{P1} over a 1:6 impedance range (e.g., varying the switching angle, α_1, of FET **2112** over approximately 0 degrees to 125 degrees) and varying phase-switched capacitor C_{P1} over a 1:10 impedance range (e.g., varying the switching angle, α_2, of FET **2110** over approximately 0 degrees to 135 degrees). Furthermore, ZTMN can be modulated to account for a ±10% variation in the load impedance Z_L (both resistive and reactive) at the frequency of operation of RF PA **2102**.

To set the correct value of switching angle, α_1, of FET **2112** and switching angle, α_2, of FET **2110** for Pi-TMN **2104** to achieve a given impedance, LUT **108** might store predetermined switching angles (e.g., α_1 and α_2) corresponding to various impedances. For example, table 3 shows an illustrative list of possible impedances Z_{TMN} that can be matched to a 50Ω load impedance Z_L and the corresponding switching angles (e.g., α_1 and α_2). The values of table 3 might be determined based upon simulation of PSIM amplifier **2100**, where FETs **2110** and **2112** are modeled having an on-state resistance of 10 mΩ and a body diode having a 0.4V forward voltage drop. The output power listed in table 3 includes power delivered at the fundamental and higher frequencies when the PSIM amplifier is supplied with a 48 VDC power supply.

TABLE 3

Switching Frequency (MHz)	α_1 (degrees)	α_2 (degrees)	TMN Impedance Z_{TMN} (Ω)	Output Power (W)
27.12	82.1	48.6	55.5 + 6.06j	19.8
27.12	64.4	68.3	125 − 1.60j	12.3
27.12	61.3	66.4	500 − 1.14j	3.5
20.86	0.10	0.10	48.9 − 1.20j	58.6
20.86	146	87.7	498 − 5.90j	5.4

As described, PSIM amplifier **2100** maintains zero-voltage-switching of all FET switches across a wide range of output power, loading impedance, and frequency of operation. For example, for illustrative PSIM amplifier **2100** to deliver an output power of 58.6 W to 50Ω load Z_L at 20.86 MHz with a power supply voltage of 48 VDC, TMN **2102** is required to provide nearly a 1:1 impedance match (e.g., Z_L=Z_{TMN}=50Ω)). Under this operating condition, the required effective shunt capacitance at nodes N_1 and N_2 is equivalent to the C_{P1} and C_{P2} capacitances, respectively, and hence FET switches **2110** and **2112** are off during the entire cycle and the drain voltage waveforms of FET switches **2110** and **2112** would be sinusoidal.

As another example, for illustrative PSIM amplifier **2100** to deliver an output power of 3.50 W to 50Ω load Z_L at 27.12 MHz with a power supply voltage of 48 VDC, TMN **2102** is required to provide an impedance Z_{TMN} of approximately 500Ω (as shown in table 3). Under this operating condition, the required effective shunt capacitance at nodes N_1 and N_2 is higher than the C_{P1} and C_{P2} capacitances, respectively, and hence FET switches **2110** and **2112** are turned on for a certain portion of the cycle while maintaining ZVS. Despite high frequency harmonic content of the drain voltage waveforms of FET switches **2110** and **2112**, the load current I_L flowing through load Z_L should remain nearly sinusoidal. Thus, PSIM amplifier **2100** is capable of providing dynamic output power control while matching into a variable load across a range of switching frequencies.

Therefore, as described herein, various embodiments provide tunable matching networks based upon phase-switched variable network reactance elements, referred to herein as phase-switched tunable matching networks (PS-TMNs). Such PS-TMNs provide rapid, high bandwidth, continuous impedance matching over a wide impedance range, while operating efficiently at high power levels without requiring high bias voltages or currents. Such PS-TMNs might be employed alone, or might also be employed in combination with other matching techniques such as discrete switched reactance banks. Described embodiments also provide zero voltage switching (ZVS) radio frequency (RF) amplifiers, referred to herein as phase-switched impedance modulation (PSIM) amplifiers. Such PSIM amplifiers might employ a PS-TMN to operate over a large frequency range by efficiently modulating output power over a wide frequency range and matching into highly variable loads (e.g., match to a wide impedance range).

Reference herein to "one embodiment" or "an embodiment" means that a particular feature, structure, or characteristic described in connection with the embodiment can be included in at least one embodiment of the claimed subject matter. The appearances of the phrase "in one embodiment" in various places in the specification are not necessarily all referring to the same embodiment, nor are separate or alternative embodiments necessarily mutually exclusive of other embodiments. The same applies to the term "implementation."

As used in this application, the words "exemplary" and "illustrative" are used herein to mean serving as an example, instance, or illustration. Any aspect or design described herein as "exemplary" or "illustrative" is not necessarily to be construed as preferred or advantageous over other aspects or designs. Rather, use of the words "exemplary" and "illustrative" is intended to present concepts in a concrete fashion.

Additionally, the term "or" is intended to mean an inclusive "or" rather than an exclusive "or". That is, unless specified otherwise, or clear from context, "X employs A or B" is intended to mean any of the natural inclusive permutations. That is, if X employs A; X employs B; or X employs both A and B, then "X employs A or B" is satisfied under any of the foregoing instances. In addition, the articles "a" and "an" as used in this application and the appended claims should generally be construed to mean "one or more" unless specified otherwise or clear from context to be directed to a singular form.

To the extent directional terms are used in the specification and claims (e.g., upper, lower, parallel, perpendicular, etc.), these terms are merely intended to assist in describing the embodiments and are not intended to limit the claims in any way. Such terms, do not require exactness (e.g., exact perpendicularity or exact parallelism, etc.), but instead it is intended that normal tolerances and ranges apply. Similarly, unless explicitly stated otherwise, each numerical value and range should be interpreted as being approximate as if the word "about", "substantially" or "approximately" preceded the value of the value or range.

Some embodiments might be implemented in the form of methods and apparatuses for practicing those methods. Further, as would be apparent to one skilled in the art, various functions of circuit elements might also be implemented as processing blocks in a software program. Described embodiments might also be implemented in the form of program code embodied in tangible media, such as magnetic recording media, hard drives, floppy diskettes, magnetic tape media, optical recording media, compact discs (CDs), digital versatile discs (DVDs), solid state memory, hybrid magnetic and solid state memory, or any other machine-readable storage medium, wherein, when the program code is loaded into and executed by a machine, such as a computer, the machine becomes an apparatus for practicing the claimed invention. Described embodiments might also be implemented in the form of program code, for example, whether stored in a storage medium, loaded into and/or executed by a machine, or transmitted over some transmission medium or carrier, such as over electrical wiring or cabling, through fiber optics, or via electromagnetic radiation, wherein, when the program code is loaded into and executed by a machine, such as a computer, the machine becomes an apparatus for practicing the claimed invention. When implemented on a processing device, the program code segments combine with the processor to provide a unique device that operates analogously to specific logic circuits. Such processing devices might include, for example, a general purpose microprocessor, a digital signal processor (DSP), a reduced instruction set computer (RISC), a complex instruction set computer (CISC), an application specific integrated circuit (ASIC), a field programmable gate array (FPGA), a programmable logic array (PLA), a microcontroller, an embedded controller, a multi-core processor, and/or others, including combinations of the above. Described embodiments might also be implemented in the form of a bitstream or other sequence of signal values electrically or optically transmitted through a medium, stored magnetic-field variations in a magnetic recording medium, etc., generated using a method and/or an apparatus as recited in the claims.

Also for purposes of this description, the terms "couple," "coupling," "coupled," "connect," "connecting," or "connected" refer to any manner known in the art or later developed in which energy is allowed to be transferred between two or more elements, and the interposition of one or more additional elements is contemplated, although not required. Conversely, the terms "directly coupled," "directly connected," etc., imply the absence of such additional elements. Signals and corresponding nodes or ports may be referred to by the same name and are interchangeable for purposes here.

It should be understood that the steps of the methods set forth herein are not necessarily required to be performed in the order described, and the order of the steps of such methods should be understood to be merely illustrative. Likewise, additional steps may be included in such methods, and certain steps may be omitted or combined, in methods consistent with various embodiments.

It will be further understood that various changes in the details, materials, and arrangements of the parts that have been described and illustrated herein might be made by those skilled in the art without departing from the scope of the following claims.

We claim:

1. A radio frequency (RF) amplifier system having an input port and an output port, the RF amplifier system comprising:

an RF amplifier having an input port coupled to the input port of the RF amplifier system and having an output port; and

a phase-switched tunable impedance network coupled between the output port of the RF amplifier and an output port of the RF amplifier system, the phase-switched tunable impedance network configured such that in response to input signal provided thereto, the phase-switched tunable impedance network varies an impedance thereof to modulate an impedance presented to the output port of the RF amplifier, wherein the impedance is modulated at a frequency related to a frequency of an RF signal at the output port of the RF amplifier.

2. The RF amplifier system of claim **1**, wherein the phase-switched tunable impedance network comprises one or more phase-switched reactive elements, each of the one or more phase-switched reactive elements configured to receive a respective control signal, and, in response to the respective control signal provided thereto, each phase-switched reactive element is provided having a corresponding desired reactance value.

3. The RF amplifier system of claim **2**, further comprising:

a controller configured to provide a respective control signal to each of the one or more phase-switched reactive elements.

4. The RF amplifier system of claim **2**, wherein each of the one or more phase-switched reactive elements comprises:

one or more reactive elements;

at least one switch;

wherein at least one of the one or more reactive elements is configured to be switched into and out of the phase-switched tunable impedance network by at least one switch associated therewith.

5. The RF amplifier system of claim **4**, wherein the at least one associated switch is operable at a switching frequency related to a frequency of an RF signal at the output port of the RF amplifier and a switching phase based upon the respective control signal.

6. The RF amplifier system of claim **4**, wherein the at least one switch is operable in a half-wave switching configuration to switch on and off once per cycle of the RF signal at the output port of the RF amplifier.

7. The RF amplifier system of claim **4**, wherein the at least one switch is operable in a full-wave switching configuration to switch on and off twice per cycle of the RF signal at the output port of the RF amplifier.

8. The RF amplifier system of claim **2**, wherein each of the one or more phase-switched reactive elements comprises a capacitor in parallel with a switch.

9. The RF amplifier system of claim **8**, wherein each of the one or more phase-switched reactive elements further comprises an inductor in series with the combination of the capacitor in parallel with the switch.

10. The RF amplifier system of claim **4**, wherein the at least one switch is operable to provide at least one of: zero voltage switching and zero current switching of said at least one switch.

11. The RF amplifier system of claim **5**, wherein the switching frequency and the switching phase are selected to provide the phase-switched reactive element having a desired reactance value.

12. The RF amplifier system of claim **11**, wherein at least one of the one or more phase-switched reactive elements is one of:

a capacitive element having a capacitive value, and wherein the capacitance value of the phase-switched capacitive element at a desired frequency is related to a physical DC capacitance value of the phase-switched capacitive element and the switching phase; and

an inductive element having an inductance value, and wherein the inductance value of the phase-switched inductive element at a desired frequency is related to a physical DC inductance value of the phase-switched inductive element and the switching phase.

13. The RF amplifier system of claim 1, wherein the impedance presented to the output port of the RF amplifier by the phase-switched tunable matching network is dynamically adapted to match a variable load impedance coupled to the output port of the RF amplifier system to an impedance of the RF amplifier.

14. The RF amplifier system of claim 13, further comprising an RF load coupled to the output port of the RF amplifier system wherein the RF load is at least one of: an antenna; a transmission line; and a plasma load.

15. The RF amplifier system of claim 4, wherein the RF amplifier comprises a switching inverter comprising at least one switching element configured to generate RF power.

16. The RF amplifier system of claim 12, wherein the controller modulates the impedance of the phase-switched tunable impedance network presented to the output port of the RF amplifier such that the RF amplifier maintains zero-voltage-switching (ZVS) of the at least one switching element of the switching inverter.

17. The RF amplifier system of claim 4, wherein switching the at least one switch of each of the one or more phase-switched reactive elements modulates an impedance transformation provided by the phase-switched tunable impedance network.

18. The RF amplifier system of claim 1 further comprising:

a filter coupled to the phase-switched tunable impedance network, the filter having a filter characteristic that reduces harmonic content generated by the phase-switched tunable impedance network that is presented to at least one of the output port of the RF amplifier and the output port of the RF amplifier system.

19. The RF amplifier system of claim 18, wherein the filter comprises one or more filter components configured to electrically isolate DC signals between the phase-switched tunable impedance network and at least one of the RF amplifier and the output port of the RF amplifier system.

20. The RF amplifier system of claim 1, wherein the phase-switched tunable impedance network comprises:

a first shunt path phase-switched variable reactive element coupled to a first node of a series path reactive element.

21. The RF amplifier system of claim 20, wherein the phase-switched tunable impedance network further comprises:

a second shunt path phase-switched variable reactive element coupled to a second node of the series path reactive element.

22. A method of operating a radio frequency (RF) amplifier system, the method comprising:

providing an RF signal to an input port of an RF amplifier;

amplifying the RF signal by the RF amplifier to provide an amplified RF signal at an output port of the RF amplifier; and

varying an impedance of a phase-switched tunable impedance network coupled between the output port of the RF amplifier and an output port of the RF amplifier system to modulate an impedance presented to the RF amplifier, wherein the impedance is modulated at a frequency related to a frequency of an RF signal at the output port of the RF amplifier.

23. The method of claim 22, wherein varying the impedance of the phase-switched tunable impedance network comprises:

receiving a control signal by the phase-switched tunable impedance network; and

in response to the control signal, electrically connecting or disconnecting at least one reactive element into or out of the phase-switched tunable impedance network at a frequency and phase related to a frequency of an RF signal at the output port of the RF amplifier to provide the phase-switched tunable impedance network having a desired reactance value.

24. The method of claim 23, wherein the phase-switched tunable impedance network comprises one or more reactive elements and at least one switch, and wherein electrically connecting or disconnecting at least one reactive element into or out of the phase-switched tunable impedance network comprises:

switching at least one of the one or more reactive elements into and out of the phase-switched tunable impedance network by at least one switch associated therewith; and

operating the at least one associated switch at a switching frequency and a switching phase based upon the respective control signal.

25. The method of claim 24, further comprising:

selecting the switching frequency and the switching phase based upon the frequency of the RF signal at the output port of the RF amplifier to provide the phase-switched reactive element having a desired reactance value.

26. The method of claim 22, further comprising:

dynamically adapting the impedance presented to the RF amplifier by the phase-switched tunable matching network to match a variable load impedance coupled to the output port of the RF amplifier system to an impedance of the RF amplifier.

27. The method of claim 22, wherein modulating the load impedance presented to the RF amplifier by the phase-switched tunable impedance network controls a power level of the amplified signal provided to the output port of the RF amplifier system.

28. The method of claim 24, wherein the amplifier comprises a switching inverter comprising at least one switching element configured to generate RF power, the method further comprising:

maintaining zero-voltage-switching (ZVS) of the at least one switching element of the switching inverter by modulating the impedance of the phase-switched tunable impedance network presented to the RF amplifier.

29. The method of claim 24, further comprising:

switching the at least one switch of the phase-switched tunable impedance network to provide at least one of: zero voltage switching and zero current switching of said at least one switch.

30. A radio frequency (RF) amplifier system having an input port and an output port, the RF amplifier system comprising:

an RF amplifier having an input port coupled to the input port of the RF amplifier system and having an output port; and

a phase-switched tunable impedance network coupled between the output port of the RF amplifier and an output port of the RF amplifier system, the phase-switched tunable impedance network configured such that in response to input signal provided thereto, the phase-switched tunable impedance network varies an impedance thereof to modulate an impedance presented to the output port of the RF amplifier,

wherein the phase-switched tunable impedance network comprises one or more phase-switched reactive elements, each phase-switched reactive element comprising at least one reactive element and at least one switch, wherein each phase-switched reactive element is configured to receive a respective control signal and operate, in response to the respective control signal provided thereto, the at least one switch at a switching frequency related to a frequency of an RF signal at the output port of the RF amplifier and a switching phase based upon the respective control signal, thereby providing a corresponding desired reactance value of each phase-switched reactive element, and wherein providing a corresponding desired reactance value for each of the one or more phase-switched reactive elements provides a desired reactance value of the phase-switched tunable impedance network, thereby modulating the impedance presented to the output port of the RF amplifier at a frequency based upon the control signals.

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