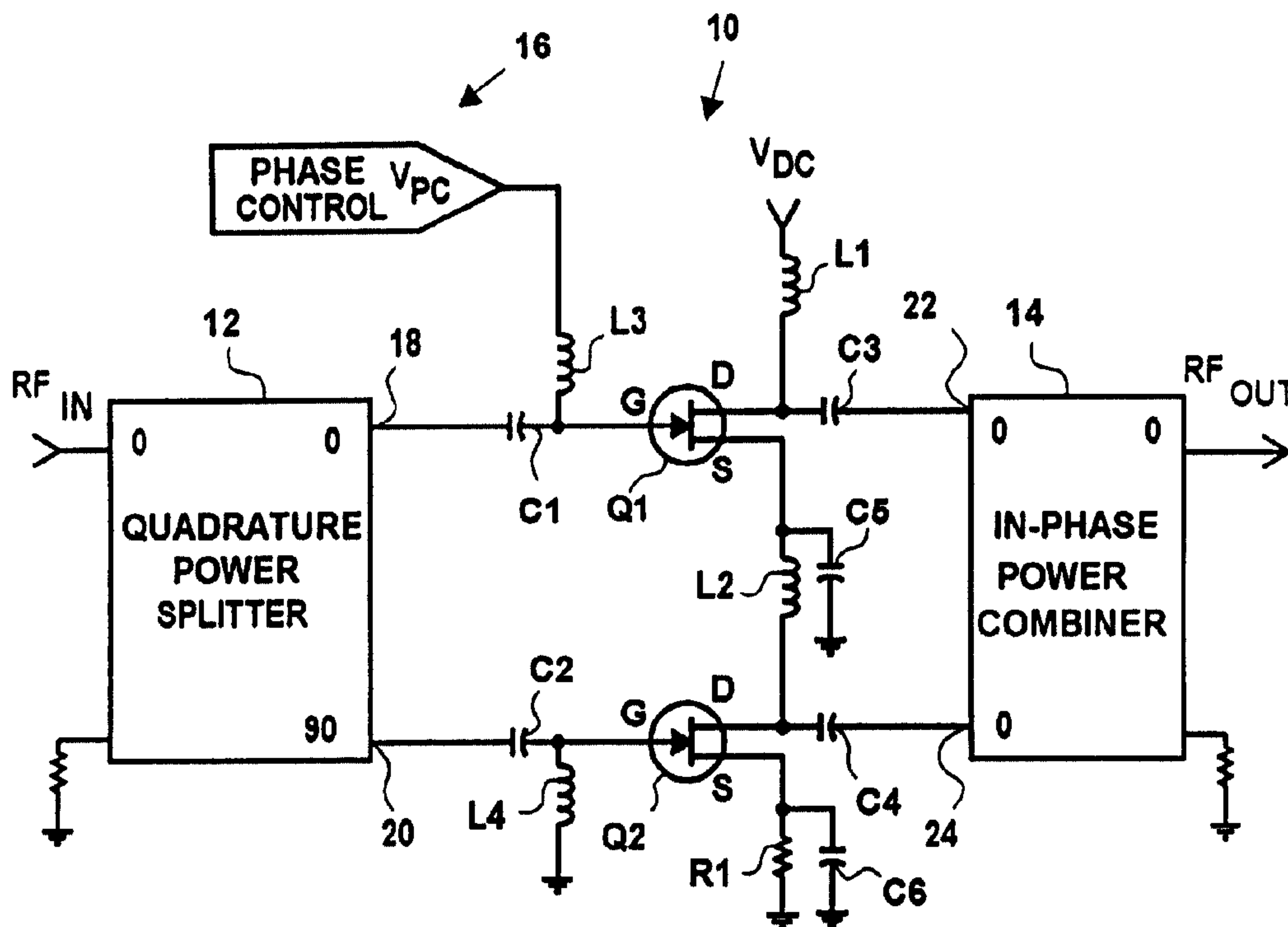




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(54) Titre : AMPLIFICATEURS DE PUISSANCE RF A DEPHASAGE VARIABLE  
(54) Title: VARIABLE PHASE-SHIFTING RF POWER AMPLIFIERS



(57) Abrégé/Abstract:

Variable phase-shifting rf power amplifiers shift rf outputs at any angle up to 90, 180, or 270 degrees, respectively, while maintaining an rf output substantially constant. The variable phase-shifting rf power amplifiers include two to four field-effect transistors that are interposed between phase splitters and combiners, and that are connected in series between a source voltage and a lower voltage. Phase shifting is achieved by selectively and variably controlling amplification of the field-effect transistors. Selective and variable control of amplification is achieved by separately and variably controlling gate voltages of the field-effect transistors, whereby a difference between the source voltage and the lower voltage is used selectively by one of the field-effect transistors and selectively proportioned between two of the field-effect transistors. Phase controls generate separate and variable phase-shifting voltages, or gate voltages, in response to a variable phase-shifting voltage.

ABSTRACT OF THE DISCLOSURE  
VARIABLE PHASE-SHIFTING RF POWER AMPLIFIERS

Variable phase-shifting rf power amplifiers shift rf outputs at any angle up to 90, 180, or 270 degrees, respectively, while maintaining an rf output substantially constant. The variable phase-shifting rf power amplifiers include two to four field-effect transistors that are interposed between phase splitters and combiners, and that are connected in series between a source voltage and a lower voltage. Phase shifting is achieved by selectively and variably controlling amplification of the field-effect transistors. Selective and variable control of amplification is achieved by separately and variably controlling gate voltages of the field-effect transistors, whereby a difference between the source voltage and the lower voltage is used selectively by one of the field-effect transistors and selectively proportioned between two of the field-effect transistors. Phase controls generate separate and variable phase-shifting voltages, or gate voltages, in response to a variable phase-shifting voltage.

TITLE OF THE INVENTION  
VARIABLE PHASE-SHIFTING RF POWER AMPLIFIERS

BACKGROUND OF THE INVENTION

Field of the Invention

5       The present invention relates generally to apparatus and method for power amplifying radio frequency (rf) or microwave rf signals. More particularly, the present invention pertains to an rf power amplifier in which the rf output is selectively phase shifted at angles up to, and beyond, 270 degrees in response to a variable phase-control voltage, or is binary-phase-shift-key modulated in response  
10to digital inputs of zero and one.

Description of the Related Art

Binary-phase-shift-key (BPSK) modulation is a form of digital modulation in which the rf carrier is phase shifted 180 degrees (inverted) as a digital input changes from 0 to 1. A demodulator, that is a part of an rf receiver, demodulates  
15these phase inversions to recover the original digital stream. Commonly, demodulation is accomplished by a Costas Loop.

A common encoder consists of the rf carrier being inserted into an rf input port of a mixer while a digital input is inserted into an input port of a local oscillator. As the digital input into the input port of the local oscillator changes  
20from an above ground voltage (1) to below ground (0), the output of the mixer changes phase from 0 degrees to 180 degrees.

If the input to the local oscillator were to change polarity (0 to 1, or 0) instantaneously, the phase of the rf output would also change polarity instantaneously. This would cause the output rf spectrum to spread to an  
25unacceptable width.

To prevent this spread in the rf output spectrum (spectrum splatter), commonly, the input to the local oscillator port is filtered (usually with a Bessel filter. As a result, the rf output decreases as the voltage to the input port of the local oscillator is decreased, and the rf output decreases to 0 when the input to the  
30local oscillator passes through 0.0 volts. Then the rf output increases in amplitude

(with inverted phase) as the voltage to the local oscillator input increases to the opposite extreme.

Therefore, as the filtered input passes through 0 volts as the polarity changes, the rf output also passes through a 0 rf output condition. This creates a problem in that the rf power amplifier section stages of conventional transmitters consists of several stages biased to Class C. In a Class C amplifier, a 0 rf input signal causes the amplifier to shut off. If a Class C amplifier were to follow the above-described encoder, it would shut off every time the input data changes state. This turning off and on of the Class C stages would cause the rf output to occupy far more of the frequency spectrum than allowed by federal regulations.

The present invention solves the above-mentioned problems with phase-shifting in general, and binary-phase-shift-key (BPSK) modulation in particular, in that the rf output stays relatively constant as the phase shifts. In one embodiment the phase shifts up to 180 degrees generally linear with a variable phase-control voltage, or shifts 180 degrees in response to a filtered BPSK input.

More particularly, the phase shifts from 0 to 90 degrees in response to a phase-control voltage increasing from 0.0 volts dc to 5.0 volts dc during which time the rf output remains substantially constant; and the rf output continues to be relatively constant as the phase shifts from 90 degrees to 180 degrees as the filtered BPSK input increases from 5.0 volts dc to 10.0 volts dc.

Since the rf output remains substantially constant during changes in the phase angle, turning off and on of Class C stages following the encoder is avoided, frequency splatter is avoided, and the occupied frequency spectrum of the rf output follows theoretical values more closely.

In the present invention, two or more solid-state amplifying devices, or FETs, are connected in series in a totem-pole arrangement, and they dividingly share, or selectably utilize, the source-voltage. That is, to phase shift the rf output to some angles the entire source-voltage is utilized by a selected one of the solid-state amplifying devices, or FETs, and to phase shift the rf output to other phase angles the source-voltage is dividingly shared by two adjacent ones of the solid-state amplifying devices.

Totem-pole arrangement of solid-state amplifying devices is taught in a paper published in *IEEE Transactions on Microwave Theory and Techniques*,

Volume 46, Number 12, of December 1998, in an article entitled, "*A 44-Ghz High IP3 InP-HBT Amplifier with Practical Current Reuse Biasing.*" As taught in this IEEE article, in totem-pole circuits two, or more, solid-state amplifying devices are used in series for dc operation, but they are used in parallel for rf operation, thereby  
5 supposedly solving the disparity between source- voltages and working voltages.

However, source-voltage sharing in the present invention is for an entirely different purpose, and functions entirely different, than that of the aforementioned IEEE article. However, the two inventions share a common problem. Unless proper rf decoupling is achieved, the maximum rf power output is extremely  
10 limited.

In addition, totem-pole, voltage-dividing, or current-sharing circuits, have been used only at low rf powers, as in the above-referenced IEEE article wherein the power was in the order of 10 milliWatts. At higher rf powers, inadequate rf decoupling has resulted in low power efficiency, oscillation, a decrease in  
15 reliability of the circuits, and destruction of the solid-state amplifying devices.

In contrast, to the extremely low rf outputs in which the prior art has been able to utilize totem-pole circuitry, apparatus and method for rf decoupling as taught herein may be used to make totem-pole circuits that are limited only by  
20 power limitations of the solid-state amplifying devices that are used in the totem pole.

More particularly, in totem-pole circuits, problems with rf decoupling are most severe between the solid-state amplifying devices. For instance, when using FETs, rf decoupling is the most critical with regard to a source terminal of any FET  
25 that is connected to a drain terminal of a next-lower FET. Capacitors and rf chokes are used for rf decoupling and rf isolating, but selection and design of capacitor decoupling is the most critical.

The next most critical location for rf decoupling is the source terminal of the lower FET when the source terminal of the lower FET is connected to an  
30 electrical ground through a resistor, as shown herein. However, if a negative bias voltage is used for the gate of the lower FET, and the source is connected directly to an electrical ground, this source terminal is already rf decoupled.

Other critical rf decoupling problems are those associated with the source-voltage to the drain of the upper FET and bias voltages to the gates of the FETs. The use of properly designed rf chokes are sufficient to provide adequate rf decoupling in these locations.

5 Unless rf decoupling is provided, as taught herein, reduced efficiency will certainly occur, and both instability and destruction of the solid-state amplifying devices are likely. This is true for both totem-pole circuitry in which a source-voltage that is excessive for a single solid-state amplifying device is dividingly shared, and for phase-shifting as taught in the present invention.

#### 10 BRIEF SUMMARY OF THE INVENTION

The present invention provides variable phase-shifting rf power amplifiers in which, in various embodiments taught herein, the rf output can be selectively shifted up to 90 degrees, up to 180 degrees, or up to 270 degrees in response to variable or preselected phase-control voltages. The resultant rf output is phase  
15 shifted without appreciably affecting the rf output power during phase-shifting, while remaining at any phase-shifted angle, or even when the rf output is digital-phase-shift-key (DPSK) modulated.

Each of the variable phase-shifting rf power amplifiers of the present invention includes a phase splitting/combining rf power amplifier and a phase  
20 control. The phase splitting/combining rf power amplifier includes 2, 3, or 4, solid-state amplifying devices, preferably FETs, for phase-shifting up to 90, 180, or 270 degrees, respectively.

The solid-state amplifying devices (FETs) in the phase splitting/combining rf power amplifier are controlled by 1, 2, or 3 phase-shifting voltages. That is, for  
25 180 degree phase-shifting, three solid-state amplifying devices are required, and two phase-shifting voltages are required. The phase controls of the present invention generate the phase-shifting voltages in response to a single phase-control voltage. However, for 90 degree phase-shifting, only one phase-shifting voltage is needed, so the phase angle increases in response to, and substantially linear to, a  
30 phase-control voltage.

In the phase splitting/combining rf power amplifiers of the present invention, an rf input is phase split into 2, 3, or 4 rf signals. The phase-split rf

signals are selectively amplified in response to phase-shifting voltages. Then the phase-split rf signals, that have been selectively amplified, are combined to provide an rf output signal whose power remains substantially constant during phase shifts up to 90 degrees, 180 degrees, 270 degrees, or more.

5 Phase splitting is accomplished by such devices as quadrature power splitters, or by a combination of 180 degree and 90 degree power splitters. Preferably, power combining is accomplished by one or more in-phase power combiners.

10 Preferably, the solid-state amplifying devices are connected in totem-pole arrangement with the solid-state amplifying devices selectively and dividingly sharing the source-voltage. Therefore, rf decoupling is critical and is accomplished as taught herein.

15 Finally, as taught herein, in designs in which the source terminal is the mounting flange of a packaged FET, as is common in high-power solid-state amplifying rf power devices, a mounting technique is used that avoids both over heating and the resultant danger of destroying the internal junctions of the solid-state amplifying device, while maintaining electrical isolation from a circuit ground.

20 In a first aspect of the present invention, a method for phase-shifting an rf output comprises: splitting an rf input into first and second rf signals that are at different phase angles; inputting the first rf signal into a first solid-state amplifying device; inputting the second rf signal into a second solid-state amplifying device; amplifying a selected one of the rf signals; and combining the rf signals subsequent to the amplifying step.

25 In a second aspect of the present invention, a method for phase-shifting an rf output comprises: splitting an rf input into first and second rf signals that are at different phase angles; inputting the first rf signal into a first solid-state amplifying device; inputting the second rf signal into a second solid-state amplifying device; proportionally amplifying the rf signals; and combining the rf signals subsequent to the amplifying step.

30 In a third aspect of the present invention, a method for binary-phase-shift-key modulating comprises: splitting an rf output into 0, 90, and 180 degree rf signals; separately amplifying the rf signals; combining the separately amplified rf

signals into a single rf output; and preventing the single rf output from decreasing to zero when the rf output is shifted 180 degrees.

In a fourth aspect of the present invention, an rf power amplifier comprises: a first solid-state amplifying device having a first higher-voltage terminal that is connected to a higher-voltage, having a first lower-voltage terminal, and  
 5 having a first control-voltage terminal; an rf choke being connected to the first lower-voltage terminal; a second solid-state amplifying device having a second higher-voltage terminal that is connected to the rf choke distal from the connection thereof to the first lower-voltage terminal, having a second lower-voltage terminal  
 10 that is connected to a lower voltage, and having a second control-voltage terminal; means, being connected between the first lower-voltage terminal and an electrical ground, for decoupling the first and second solid-state amplifying devices; and the means for decoupling comprises an rf effective series resistance of less than 0.4 divided by an rf output, in Watts, of the rf power amplifier.

#### 15 BRIEF DESCRIPTION OF THE SEVERAL VIEWS OF THE DRAWINGS

FIGURE 1 is a variable phase-shifting rf power amplifier of the present invention in which two, n-channel, gallium arsenide FETs are stacked to selectively utilize a source-voltage, and in which an rf output can be shifted up to 90 degrees proportional to, and substantially linearly with, a single phase-control voltage;

20 FIGURE 2 is a phase splitting/combining rf power amplifier, in which three FETs are stacked to selectively utilize a source-voltage, that when combined with a phase control of FIGURE 3, becomes a variable phase-shifting rf power amplifier in which a phase angle of an rf output can be shifted up to 180 degrees in response to two phase-shifting voltages;

25 FIGURE 3 is a phase control, that generates two phase-shifting voltages in response to a variable phase-control voltage, and that when combined with the phase splitter/combiner rf power amplifier of FIGURE 2, becomes a variable phase-shifting rf power amplifier in which the rf output can be phase shifted up to 180 degrees substantially linear with a phase-control voltage;

30 FIGURE 4 is a phase splitting/combining rf power amplifier, in which four FETs are stacked to selectively utilize a source-voltage, that when combined with a phase control of FIGURE 5, becomes a variable phase-shifting rf power amplifier in

which a phase angle of an rf output can be shifted up to 270 degrees in response to three phase-shifting voltages;

FIGURE 5 is a phase control, that generates three phase-shifting voltages in response to a single phase-control voltage, and that when combined with the phase  
5 splitter/combiner rf power amplifier of FIGURE 4, becomes a variable phase-shifting rf power amplifier in which the rf output can be phase shifted up to 270 degrees substantially linear with a single phase-control voltage;

FIGURE 6 is a model for simulating a microwave inductor;

FIGURE 7 is model for simulating a microwave capacitor;

10 FIGURE 8 shows the use of multiple decoupling capacitors to minimize the equivalent series resistance (ESR) of the decoupling capacitors; and

FIGURE 9 is a side elevation, in partial cross section, of a high-power rf FET that is mounted to achieve maximum thermal conduction while maintaining electrical isolation of the source-terminal from an electrical ground.

## 15 DETAILED DESCRIPTION OF THE INVENTION

Referring now to FIGURE 1, a variable phase-shifting rf power amplifier 10 includes solid-state amplifying devices, field-effect transistors, or FETs, Q1 and Q2 that are connected in series between a higher-voltage, or source-voltage  $V_{DC}$  and a lower voltage or a ground. That is, a first rf choke L1 connects the source-voltage  
20  $V_{DC}$  to a drain terminal of the FET Q1, a second rf choke L2 connects a source terminal of the FET Q1 to a drain terminal of the FET Q2, and a resistor R1 connects a source terminal of the FET Q2 to a ground.

The variable phase-shifting rf power amplifier 10 also includes an rf quadrature power splitter 12 and an rf in-phase power combiner 14. The  
25 quadrature power splitter 12 is connected to gate terminals of the FETs Q1 and Q2, respectively, by coupling capacitors C1 and C2. The rf power combiner 14 is connected to drain terminals of the FETs Q1 and Q2, respectively, by coupling capacitors C3 and C4. And source terminals of the FETs Q1 and Q2 are connected to an electrical ground by decoupling capacitors C5 and C6, respectively.

30 A phase control 16 provides a phase-shifting voltage  $V_{PS}$  in response to a phase-control voltage  $V_{PC}$ , and supplies the phase-shifting voltage  $V_{PS}$  to the gate terminal of the FET Q1 through a third rf choke L3 as a variable bias voltage. The

resistor R1 supplies a negative gate-to-source bias for the gate terminal of the FET Q2 through a fourth rf choke L4. The resistor R1, in setting the gate-to-source bias for the FET Q2, controls current flow through the FETs, Q1 and Q2, thereby controlling rf power amplification of the variable phase-shifting rf power amplifier  
5 10.

In operation, an rf input signal  $RF_{IN}$  of the variable phase-shifting rf power amplifier 10 is split in the rf power splitter 12, selectively amplified in the FET Q1 and/or in the FET Q2, and combined in the rf power combiner 14 to provide a power amplified output at an rf output terminal  $RF_{OUT}$  that is selectively phase  
10 shifted.

The amplifying function of the FETs Q1 and Q2 is maintained by using rf chokes L1, L2, L3, and L4, to keep the rf signal from coupling onto the dc bias lines and to prevent rf interference between FETs Q1 and Q2; and decoupling capacitors, C5 and C6, are used to keep the source terminals of both FETs, Q1 and  
15 Q2, at an rf ground.

As taught herein, the performance of rf power amplifiers that series connect FETs, or other solid-state amplifying devices, rests heavily on correct design and application of rf chokes, such as the rf chokes L1, L2, L3, and L4 of FIGURE 1, and decoupling capacitors, such as the decoupling capacitors C5 and  
20 C6 of FIGURE 1. Therefore, rf choke and decoupling capacitor design will be considered in greater detail after considering various other embodiments of the present invention.

The voltage to the drain terminal D of the upper FET Q1 cannot exceed the specified FET drain-to-source voltage ( $V_{ds}$ ). Or, if the FET Q1 were replaced  
25 by a bipolar transistor, not shown, the collector-to-emitter voltage ( $V_{ce}$ ) could not exceed specifications. Therefore, in the case of GaAsFETs the source- voltage should be 12 volts dc ( $V_{ds} + V_{pinchoff}$  of the lower FET Q2).

In operation, if the phase-shifting voltage,  $V_{PS}$  is lowered to 0.0 volts dc by the phase control 16, 10.0 volts dc will be applied across the FET Q1, and 0.0  
30 volts dc will be applied across the FET Q2. Since the gain of FETs, such as the FETs Q1 and Q2, is approximately a linear function of the drain-to-source voltage, an rf output of the FET Q1 will be at maximum gain while an rf output of the FET Q2 will be at minimum gain.

At this time, the rf in-phase power combiner 14 will output half of the rf power to the rf output terminal  $RF_{OUT}$  and half of the rf power to the internal or external load. More importantly, the half delivered to rf output terminal  $RF_{OUT}$  will be in-phase with a first rf signal at an upper rf output terminal 18 of the quadrature power splitter 12, that is disregarding inversion of the FET Q1.

If the phase-shifting voltage is now raised to 10.0 volts dc by the phase control 16, 0.0 volts dc will be applied across the FET Q1, and 10.0 volts dc will be applied across the FET Q2. The FET Q1 will now be at a minimum gain while the FET Q2 will be at maximum gain. In this case, the output of the in-phase rf power combiner 14 will be in-phase with a second rf signal at a lower rf output terminal 20 of the quadrature power splitter 12. That is, the phase will have been shifted 90 degrees. Again, half of the power is delivered to the rf output terminal  $RF_{OUT}$  and half is delivered to the internal or external load.

If the phase-shifting voltage is set to 5.0 volts dc by the phase control 16, 5.0 volts dc will be applied across both the FET Q1 and the FET Q2, and both FETs will operate at half gain. In this case, an upper rf input terminal 22 and a lower rf input terminal 24 to the rf in-phase power combiner 14 will be equal in amplitude but 90 degrees out of phase.

At this time, the rf output terminal  $RF_{OUT}$  of the rf in-phase power combiner 14 remains at half power but is 45 degrees out of phase with the upper rf input terminal 22. As before, half of the power will be delivered to the internal or external load.

Thus, it can be seen that phase control 16 is effective to shift the phase of the variable phase-shifting rf power amplifier 10 monotonically, and with reasonable linearly, from 0 to 90 degrees as the phase-control voltage is varied from 0.0 volts dc to 10.0 volts dc.

Finally with regard to FIGURE 1, alternately, instead of the quadrature power splitter 12 and the rf in-phase power combiner 14, an in-phase splitter and a quadrature combiner may be used.

Referring now to FIGURES 2 and 3, a variable phase-shifting rf power amplifier 30 includes both a phase splitting/combining rf power amplifier 32 of FIGURE 2 and a phase control 34 of FIGURE 3. The variable phase-shifting rf

power amplifier 30 has a phase-shift range of 180 degrees, which is twice that of the variable phase-shifting rf power amplifier 10 of FIGURE 1.

The phase splitting/combining rf power amplifier 32 of FIGURE 2 includes a 180 degree power splitter 36, a 90 degree power splitter 38A, input terminals  
5 that shift an rf output in response to phase-shifting voltages,  $V_{PS1}$  and  $V_{PS2}$ , solid-state amplifying devices, field-effect transistors, or FETs, Q1, Q2, and Q3, and 0 degree power combiners, 40A and 40B.

Also, the phase splitting/combining rf power amplifier 32 includes coupling capacitors C1, C2, C3, and C4, decoupling capacitors C5 and C6, rf  
10 chokes L1, L2, L3, and L4, and a resistor R1 as shown in FIGURE 1. In addition, the phase splitting/combining rf power amplifier 32 includes coupling capacitors C7 and C8, decoupling capacitor C9, and rf chokes L5 and L6.

If phase-shifting voltages,  $V_{PS1}$  and  $V_{PS2}$  are at 0.0 volts dc, 10.0 volts dc will be applied across the FET Q1 and 0.0 volts dc will be applied across the FETs  
15 Q2 and Q3. At this time, since the gain of the FETs Q1, Q2, and Q3 is approximately a linear function of the applied voltage from drain to source, the FET Q1 will be at maximum gain while the FETs Q2 and Q3 will be at minimum gain, and the rf output ( $RF_{OUT}$ ) will be at 0 degrees relative to the rf input signal ( $RF_{IN}$ ), that is disregarding inversion of the FET Q1.

20 If the phase-shifting voltage  $V_{PS1}$  is raised to 10.0 volts dc, and phase-shifting voltages  $V_{PS2}$  and  $V_{PS3}$  remain at 0.0 volts dc, 10.0 volts dc will be applied across the FET Q2, and 0.0 volts dc will be applied across the FETs Q1 and Q3. The FET Q2 will now be at maximum gain while the FETs Q1 and Q3 will be at minimum gain. In this case, the rf output ( $RF_{OUT}$ ) will be at 90 degrees relative to  
25 the rf input signal ( $RF_{IN}$ ). Again, this disregards inversion of the FET Q2.

Similarly to FIGURE 1, if the phase-shifting voltage  $V_{PS1}$  is 5.0 volts dc, and the phase-shifting voltage  $V_{PS2}$  is at 0.0 volts dc, the rf output ( $RF_{OUT}$ ) will be at 45 degrees relative to the rf input signal ( $RF_{IN}$ ). By proper application of the phase-shifting voltages,  $V_{PS1}$  and  $V_{PS2}$ , the phase angle of the variable phase-shifting rf  
30 power amplifier 30 can be made to vary monotonically and reasonably linearly from 0 degrees to 180 degrees.

Referring again to FIGURES 2 and 3, as noted above, the variable phase-shifting rf power amplifier 30 includes both the phase splitting/combining rf power

amplifier 32 of FIGURE 2 and the phase control 34 of FIGURE 3. The phase control 34 generates phase-shifting voltages  $V_{PS1}$  and  $V_{PS2}$  for use by the phase splitting/combining rf power amplifier 32. These phase-shifting voltages,  $V_{PS1}$  and  $V_{PS2}$ , are generated in response to the phase-control voltages  $V_{PC}$  that are  
 5 adjustably, or selectably, applied to the phase control 34 of FIGURE 3.

The phase control 34 of FIGURE 3 includes amplifiers U1 and U2 which are rail-to-rail operational amplifiers. In addition, the phase control 34 includes resistors R2, R3, R4, and R5 that set the gain of the amplifiers, U1 and U2, and that set the voltage at which the amplifier U2 starts amplifying.

10 The amplifier U1 is biased to start amplifying at the phase-control voltage  $V_{PC}$  of 0.0 volts, and the amplifier U2 is biased to start amplifying at the phase-control voltage  $V_{PC}$  of 5.0 volts. In the schematic shown in FIGURE 3, the resistors R2, R3, R4, and R5 all have the same resistances, which, for instance, may have resistances of 10K ohms.

15 In response to the phase-control voltage  $V_{PC}$  of 0.0 volts, the phase control 34 produces phase-shifting voltages  $V_{PS1}$  and  $V_{PS2}$  of 0.0 volts, dc. In response to increases in the phase-control voltage  $V_{PC}$ , the phase-shifting voltage  $V_{PS1}$  increases to 5.0 volts while keeping the phase-shifting voltage  $V_{PS2}$  at 0.0 volts dc. Phase-control voltages  $V_{PC}$  of 0.0, 2.5, 5.0, 7.5, and 10.0 volts produce phase angles of 0,  
 20 45, 90, 135, and 180 degrees, respectively.

With further increases in the phase-control voltage  $V_{PC}$ , when the phase-shifting voltage  $V_{PS1}$  reaches 10.0 volts dc, it remains at this level while the phase-shifting voltage  $V_{PS2}$  increases from 0.0 volts to 10.0 volts dc.

25 Thus, it can be seen that by combining the phase control 34 with the phase splitting/combining rf power amplifier 32, the resultant variable phase-shifting rf power amplifier 30 can be phase shifted monotonically and reasonably linearly from 0 degrees to 180 degrees as the phase-control voltage  $V_{PC}$  is increased.

30 Referring now to FIGURES 4 and 5, a variable phase-shifting rf power amplifier 50 includes both a phase splitting/combining rf power amplifier 52 of FIGURE 4 that requires phase-shifting voltages  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$ , and a phase control 54 of FIGURE 5 that generates the phase-shifting voltages  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$  in response to the adjustable or selectable phase-control voltage  $V_{PC}$ . The

variable phase-shifting rf power amplifier 50 has a phase-shift range of 270 degrees, as opposed to 180 degrees for the variable phase-shifting rf power amplifier 30 of FIGURES 2 and 3.

Referring now to FIGURE 4, the phase splitting/combining rf power amplifier 52 includes the 180 degree power splitter 36, the 90 degree power splitter 38A, a 90 degree power splitter 38B, input terminals that accept phase-shifting voltages  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$ , solid-state amplifying devices, field-effect transistors, or FETs, Q1, Q2, Q3, and Q4, the 0 degree power combiners 40A and 40B, and an other 0 degree power combiner 40C.

The phase splitting/combining rf power amplifier 52 includes coupling capacitors, decoupling capacitors, and rf chokes as shown in FIGURES 1 and 2, and as named in conjunction therewith. In addition, the phase-splitting/combining rf power amplifier 52 includes coupling capacitors C10 and C11, decoupling capacitor C12, and rf chokes L7 and L8.

If phase-shifting voltages  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$ , are all at 0.0 volts dc, 10.0 volts dc will be applied across the FET Q1 and 0.0 volts dc will be applied across the FETs Q2, Q3, and Q4. Since the gain of the FETs, Q1, Q2, Q3, and Q4 is approximately a linear function of the applied voltage from drain to source, the FET Q1 will be at maximum gain while the FETs Q2, Q3, and Q4 will be at minimum gain. The rf output ( $RF_{OUT}$ ) will then be at 0 degrees relative to the rf input signal ( $RF_{IN}$ ), that is disregarding inversion of the FET Q1.

If the phase-shifting voltage  $V_{PS1}$  is now raised to 10.0 volts dc and the phase-shifting voltages  $V_{PS2}$  and  $V_{PS3}$  remain at 0.0 volts dc, 10.0 volts dc will be applied across the FET Q2, and 0.0 volts dc will be applied across the FETs Q1, Q3, and Q4. The FET Q2 will now be at maximum gain while the FETs Q1, Q3, and Q4 will be at minimum gain. In this case, the rf output ( $RF_{OUT}$ ) will be at 90 degrees relative to the rf input signal ( $RF_{IN}$ ), again disregarding inversion of the FET Q2.

Similarly to FIGURE 1, if the phase-shifting voltage  $V_{PS1}$  is at 5.0 volts dc, and the phase-shifting voltages  $V_{PS2}$  and  $V_{PS3}$  are at 0.0 volts dc, the rf output ( $RF_{OUT}$ ) will be at 45 degrees relative to the rf input signal ( $RF_{IN}$ ). By proper application of the phase-shifting voltages  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$ , the phase of the

phase-shifting rf power amplifier 50 can be made to vary monotonically and reasonably linearly from 0 degrees to 270 degrees.

These concepts can be extended to even wider phase control by applying the principles set forth in conjunction with FIGURE 4. Optionally, the splitters and  
5 combiners can be at phase angles other than 0 degrees, 90 degrees, and 180 degrees.

Referring again to FIGURES 4 and 5, as noted above, the variable phase-shifting rf power amplifier 30 includes both the phase splitting/combining rf power amplifier 52 of FIGURE 4 and the phase control 54 of FIGURE 5. The phase  
10 control 54 generates phase-shifting voltages  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$  for use by the phase splitting/combining rf power amplifier 52 in response to the phase-control voltage  $V_{PC}$  that is adjustably or selectably applied to the phase control 54 of FIGURE 5.

The phase control 54 of FIGURE 5 includes amplifiers U1, U2, and U3  
15 which are rail-to-rail operational amplifiers. In addition, the phase control 54 includes resistors R6, R7, R8, R9, R10, R11, R12, and R13 that set the gain of the amplifiers, U1, U2, and U3, to be 4.0. Resistances of the resistors R6, R7, R8, R9, R10, R11, R12, and R13, preferably are 30K, 10K, 30K, 30K, 15K, 30K, 15K and 30K, respectively, but all may be at resistances that are any reasonable multiple or  
20 fraction thereof.

The amplifiers, U1, U2, and U3, are biased to start amplifying at different phase-control voltages  $V_{PC}$  of 0.0, 2.5, 5.0, and 7.5 volts by resistances as listed above; so that phase-control voltages  $V_{PC}$  of 0.0, 2.5, 5.0, 7.5, and 10.0 volts produce phase angles of 0, 45, 90, 135, and 180 degrees, respectively.

25 More particularly, in response to the phase-control voltage  $V_{PC}$  of 0.0 volts, the phase control 34 produces phase-shifting voltages,  $V_{PS1}$ ,  $V_{PS2}$ , and  $V_{PS3}$ , of 0.0 volts, dc. In response to increases in the phase-control voltage  $V_{PC}$ , the phase-shifting voltage  $V_{PS1}$  increases to 10.0 volts while keeping the phase-shifting voltage  $V_{PS2}$  at 0.0 volts dc.

30 With further increases in the phase-control voltage  $V_{PC}$ , when the phase-shifting voltage  $V_{PS1}$  reaches 10.0 volts dc, it remains at this level while the phase-shifting voltage  $V_{PS2}$  increases from 0.0 volts to 10.0 volts dc. In like manner, after the phase-shifting voltages,  $V_{PS1}$  and  $V_{PS2}$ , both reach 10.0 volts dc, they remain at

10.0 volts dc while additional increases in the phase-control voltage  $V_{PC}$  increase the phase-shifting voltage  $V_{PS3}$  from 0.0 to 10.0 volts dc.

Thus, combining the phase splitting/combining rf power amplifier 52 with the phase control 54 provides the variable phase-shifting rf power amplifier 50 in  
5 which the rf output can be phase shifted monotonically and reasonably linearly from 0 degrees to 270 degrees as the phase-control voltage  $V_{PC}$  is increased.

Referring again to FIGURES 1, 2, and 4, as stated previously, the amplification function of the FETs, such as the FETs Q1 and Q2, is maintained by using rf chokes, such as the rf chokes L1, L2, L3, and L4, to keep the rf signal from  
10 getting onto the dc bias lines and to prevent rf interference between the series-connected FETs; and decoupling capacitors, such as the decoupling capacitors C5 and C6, are used to keep the sources of FETs at an rf ground.

The selection of the decoupling capacitors and rf chokes are both critical to the rf performance of the circuits, particularly for high-power rf amplifiers,  
15 although selection of decoupling capacitors is the most critical. Decoupling capacitors, such as the decoupling capacitors C5, C6, C9, and C12 are selected for both resonant frequencies at or very near to the circuit operating frequency and the lowest possible effective (or equivalent) series resistances (ESRs).

The rf chokes, such as the rf chokes L1, L2, L3, L4, L5, L6, L7, and L8  
20 preferably are inductors with self-resonant frequencies at or very near to the circuit operating frequency.

Referring now to FIGURE 6, the microwave circuit model of an inductor is a series resistor  $R_s$  and inductor  $L$  in parallel with a capacitor  $C$ . The resistor  $R_s$  represents the dc coil resistance along with the increased wire resistance at rf  
25 frequencies due to the skin effect (the effect of the current being concentrated nearer to the surface of the wire) as the operational frequency is increased. The capacitor  $C$  represents the distributed capacitance between the parallel windings of the coils. Inductance of the inductor  $L$  is the nominal component inductance.

At operation below the self-resonant frequency, the impedance of an  
30 inductor increases as frequency increases. At the inductor self-resonant frequency, the inductor, as represented by a parallel L/C circuit of FIGURE 6, resonates as an open circuit creating a maximum impedance to the rf signal. At operation higher than the self-resonant frequency, the distributed capacitance of the capacitor  $C$

dominates the rf impedance resulting in the impedance decreasing with increasing frequency. The inductor self-resonant frequency is given as:  $F_{SR} = 1/[2\pi*\sqrt{LC}]$ .

The resistance of the series resistor  $R_s$  limits the maximum impedance of the self-resonant inductor. That is, the quality factor (Q) of the inductor is the ratio  
5 of an inductor's reactance to its series resistance. High-Q inductors, with very low resistances, have very high self-resonant impedances, but for only a narrow bandwidth. Lower-Q inductors, with higher resistances, have lower self-resonant impedances for a much broader bandwidth.

This self-resonant feature is used in the circuit to prevent the rf signal from  
10 coupling onto the dc bias lines and to aide the decoupling capacitors in preventing rf crosstalk between the two, or more, FETs. For narrow-band operation, very high-Q inductors are desired to maximize series impedance. Quarter wave transformers may also be used for this function in narrow-band applications. For broad-band operation, lower-Q inductors are desired to obtain a high impedance across a  
15 larger bandwidth. In either application, the inductor must be capable of passing the maximum dc current without breakdown.

Utilizing the self-resonant characteristics of decoupling capacitors, such as the decoupling capacitors C5, C6, C9, and C12, is required to optimize rf performance while maximizing dc-rf conversion efficiency, particularly in  
20 applications where the rf power exceeds 100 milliWatts.

Referring now to FIGURE 7, the microwave circuit model of a capacitor is an inductor L in series with a resistor  $R_s$  in series with a capacitor C. The inductor L represents the inductance of the leads and the capacitor plates. The resistor  $R_s$  represents the equivalent series resistance, or ESR, of the capacitor. Capacitor  
25 dielectric losses, metal plate losses, and skin effects all contribute to the ESR. The capacitor C is the nominal component capacitance.

These parasitic effects of a capacitor at microwave frequencies alter its impedance characteristics in the opposite manner as that of an inductor. At operation below the self-resonant frequency, a capacitor decreases in impedance as  
30 frequency increases. At the capacitor self-resonant frequency, a capacitor, as represented by a series L/C circuit of FIGURE 7, resonates as a short circuit creating a minimum impedance to the rf signal. At frequencies higher than the self-resonant frequency, the lead and plate inductance L dominates the rf

impedance resulting in the impedance increasing with increasing frequency. The capacitor self-resonant frequency equation is:  $F_{SR} = 1/[2\pi*\sqrt{LC}]$ , which is the same as for the inductor.

The rf impedance of a capacitor at self-resonant frequency is equal to the ESR. As in the case of the inductor L, Q of a capacitor is the ratio of a capacitor's reactance to its ESR, or alternatively Q is 1/DF where DF is the dissipation factor of the capacitor. High-Q capacitors, with very low ESR, have very low self-resonant impedances, but for only a narrow bandwidth. Lower-Q capacitors, with higher ESR, have lower self-resonant impedances for a much broader bandwidth.

Presently, the preferred capacitor dielectric to minimize capacitor ESR is porcelain. Porcelain has a dissipation factor (DF) of 0.00007, the lowest of all currently available capacitor dielectrics.

To minimize the rf impedance from the FET source terminal to a circuit ground, decoupling capacitors with self-resonant frequencies at or very near to the amplifier operational frequency are required in higher rf power applications.

The power dissipated in the decoupling capacitor is  $P_{DISS} = I^2*ESR$ , where I is the root-mean-square, or rms, of the rf current through the capacitor. Alternatively,  $P_{DISS} = P_{RF}*ESR/Z$  where Z is the circuit load impedance, typically 50 ohms, and  $P_{RF}$  is the rf output power of the FET.

For optimal performance, the ratio of FET rf output power  $P_{RF}$  to decoupling capacitor power dissipated  $P_{DISS}$ , should be no less than 2000 for medium rf power, which is defined as 100 milliWatts to 2.0 Watts FET rf output power. For high-power rf applications, which is defined as FET output power greater than 2.0 Watts, the  $P_{RF}/P_{DISS}$  ratio should be no less than 5000.

Very high-Q decoupling capacitors are necessary to minimize series impedance to a circuit ground, whether it be for narrow-band, or wide-band operation. For broad-band operation, multiple high-Q decoupling capacitors with self-resonant frequencies selected at several points in the operating frequency band are optimally selected for minimum ESR across a broad frequency band.

Referring now to FIGURE 8, two or more multiple porcelain dielectric capacitors, each with self-resonant frequencies at or near the amplifier operational frequency, are connected in parallel from the FET source terminal to a circuit

ground to achieve the low required decoupling capacitor ESR for high power rf applications.

Paralleling a plurality of capacitors at the self-resonant frequency divides the ESR in the same manner as paralleling resistors. However, if a capacitor is not  
 5 available with a resonant frequency that closely matches an operating frequency for narrow-band operation, two paralleled capacitors are chosen with one having a resonant frequency above the narrow-band frequency, and the other having a resonant frequency below the narrow-band frequency.

Referring now to FIGURES 1, 2, and 4, preferably the effective series  
 10 resistances of the decoupling capacitors C5, C6, C9, and/or C12 each have an effective series resistance of less than 0.4 ohms divided by the rf output power. More preferably, all of these decoupling capacitors have an effective series resistance of 0.20 ohms divided by the rf output power.

If the required ESR, as calculated by either of the formulas given above, for  
 15 any or all of the decoupling capacitors C5, C6, C9, and/or C12 cannot be met by a single capacitor, any or all may be replaced by any number of paralleled capacitors Ca-n, as shown in FIGURE 8.

Porcelain capacitors presently have the lowest dielectric resistance and are preferred for minimizing the effective rf impedance. Porcelain capacitors, model  
 20 600S, manufactured by *American Technical Ceramics* of Huntington Station, New York, are suitable for rf decoupling as taught herein.

Model 600S capacitors that are available from American Technical Ceramics, their self resonant frequencies, their capacities, and their effective series resistances, are included in the following table.

25

Table 1: Porcelain Capacitors  
 Self Resonant Frequencies vs. ESRs

	Self Resonant Freq.	Capacitance	ESR
	1 Ghz	100 pF	0.07 ohms
	2 Ghz	40 pF	0.09 ohms
30	4 Ghz	15 pF	0.15 ohms
	8 Ghz	3 pF	0.20 ohms
	16 Ghz	1 pF	0.30 ohms

Referring now to FIGURE 8 and Table 1, as an example of capacitor paralleling to achieve a required ESR, assume an rf output of 5.0 Watts, using the 0.2 ohms/Watts criteria, the ESR of the decoupling capacitor should be 0.04 ohms. Assuming an operating frequency of 4.0 Ghz, from Table 1, the ESR for a porcelain  
5 capacitor is 0.15 ohms, so four capacitors must be paralleled to achieve the required ESR.

Packaged FETs typically have a considerable source lead parasitic inductance. By choosing a decoupling capacitor, or capacitors, with a value that resonates with the source lead inductance, the true FET source impedance to a  
10 circuit ground is further reduced.

Therefore, the package, or lead, inductance of the capacitor, or capacitors, should be considered in the equation for resonance when selecting a capacitor to resonate with the FET source lead inductance. Additionally, several parallel capacitors with a combined reactance that resonates with the FET source lead  
15 inductance are selected to minimize the decoupling capacitor ESR and maximize efficiency in high-power rf applications (FET rf output in excess of 2.0 Watts).

Often in high-power packaged FETs the source terminal is the body of the device and is connected to a mounting flange. Conventionally, the flange is connected directly to a circuit ground with metallic screws to achieve minimal rf  
20 impedance to an electrical ground and to maximize thermal conductivity between the FET and a circuit ground, which is most often a chassis serving as a heat sink to the FET. However, in the present invention, the source terminals of the FETs are electrically isolated from a circuit ground.

Referring now to FIGURE 9, a thermally conductive, electrically insulating  
25 pad 60 is inserted between a FET mounting flange 62 of a FET 64 and a heat sink, or chassis, 66 to allow the dissipated heat of the FET 64 to flow from the FET 64 to the heat sink 66 while maintaining electrical isolation. The electrical insulating material of the pad 60 should have no higher than 0.5 °C/Watt thermal resistance. An insulating material with a higher thermal resistance, combined with the thermal  
30 resistance of the FET and the ambient temperature, may result in the internal junction temperature of the FET being excessive, thereby causing reduced reliability or destruction of the FET.

A suitable material for the insulating material is *DeltaPad Thermally Conductive Insulator*, Part Number 174-9 Series, manufactured by *Wakefield Engineering* of Pelham, NH. The material for the insulating pad 60 is 0.22 millimeters (0.009 inches) thick, has a thermal resistance of  $0.25^{\circ} \text{C/W}$ , a resistivity of  $10^{13}$  megohms per cubic centimeter of volume, and a 5000 volt breakdown.

The mounting flange 62 is held in heat-conducting contact with the insulating pad 60 and with the heat sink 66, with non-ferrous, or non-conductive, screws 68. The tensile strength and stretching of the screw material along with the manufacturer-recommended FET mounting torque must be taken into account when selecting fasteners.

Although the preceding discussion has focused on use of FETs, bipolar silicon transistors, and other solid-state amplifying devices may be used. However, FETs are preferred because of their high gain, thereby reducing the total number of amplification stages that are required to achieve the desired rf power output. Therefore, it should be understood that the principles taught herein may be applied to other types of solid-state amplifying devices.

In summary, the present invention can be characterized as phase splitting an rf input into rf signals that are at different phase angles, selectively amplifying one and an other of the rf signals, and combining the rf signals subsequent to the amplifying step.

The present invention can be characterized as phase splitting an rf input into rf signals that are at different phase angles, amplifying the rf signals at selective proportions, and combining the rf signals subsequent to the amplifying step.

The present invention can be characterized as applying a voltage across two FETs that are connected in series, and selectively utilizing the voltage in one or an other of the FETs.

The present invention also can be characterized as applying a voltage across two FETs that are connected in series, and selectively proportioning the voltage between the FETs.

The present invention can be characterized as phase-shifting an rf output up to 180, 270, or more, degrees without the rf output decreasing to zero, or even changing the rf output appreciably.

Finally, the present invention can be characterized as providing optimum rf decoupling, especially by reducing the effective series resistance (ESR) of decoupling capacitors, thereby removing power limitations from rf power amplifiers in which solid-state amplifying devices, such as FETs, are connected in series between a source-voltage and a lower-voltage.

While specific apparatus and method have been disclosed in the preceding description, it should be understood that these specifics have been given for the purpose of disclosing the principles of the present invention, and that many variations thereof will become apparent to those who are versed in the art.

Therefore, the scope of the present invention is to be determined by claims included herein without any limitation by numbers that may be parenthetically inserted in the claims.

#### INDUSTRIAL APPLICABILITY

The adjustable phase-shifting power amplifiers of the present invention are applicable to applications wherein varactors have used to variably shift the phase of an rf signal, and especially in applications wherein significant rf power is involved, since varactors are extremely limited in their power-handling capacity and their outputs require considerable power amplification.

## CLAIMS

The embodiments of the invention in which exclusive property or privilege is claimed are defined as follows:

What is claimed is:

1. A method for phase-shifting an rf output which comprises:
  - a) splitting an rf input into first and second rf signals that are at different phase angles;
  - b) inputting said first rf signal into a first solid-state amplifying device;
  - c) inputting said second rf signal into a second solid-state amplifying device;
  - d) amplifying a selected one of said rf signals; and
  - e) combining said rf signals subsequent to said amplifying step.
  
2. A method as claimed in Claim 1 in which said amplifying of said selected one of said rf signals comprises biasing one of said solid-state amplifying devices.
  
3. A method as claimed in Claim 1 in which:
  - a) said method further comprises connecting said solid-state amplifying devices in series between a source voltage and a lower voltage; and
  - b) said amplifying of said selected one of said rf signals comprises utilizing a difference in said voltages entirely in said one solid-state amplifying device.
  
4. A method for phase-shifting an rf output which comprises:
  - a) splitting an rf input into first and second rf signals that are at different phase angles;
  - b) inputting said first rf signal into a first solid-state amplifying device;
  - c) inputting said second rf signal into a second solid-state amplifying device;
  - d) proportionally amplifying said rf signals; and
  - e) combining said rf signals subsequent to said amplifying step.

5. A method as claimed in Claim 4 in which said step of proportionally amplifying comprises proportionally biasing said solid-state amplifying devices.

6. A method as claimed in Claim 4 in which:

a) said method further comprises connecting said solid-state amplifying devices in series between a source voltage and a lower voltage; and

b) said step of proportionally amplifying comprises proportionally dividing a difference between said voltages between said solid-state amplifying devices.

7. A method for binary-phase-shift-key modulating which comprises:

a) splitting an rf output into 0, 90, and 180 degree rf signals;

b) separately amplifying said rf signals;

c) combining said separately amplified rf signals into a single rf output;

and

d) preventing said single rf output from decreasing to zero when said rf output is shifted 180 degrees.

8. An rf power amplifier which comprises:

a first solid-state amplifying device having a first higher-voltage terminal that is connected to a higher-voltage, having a first lower-voltage terminal, and having a first control-voltage terminal;

an rf choke being connected to said first lower-voltage terminal;

a second solid-state amplifying device having a second higher-voltage terminal that is connected to said rf choke distal from said connection thereof to said first lower-voltage terminal, having a second lower-voltage terminal that is connected to a lower voltage, and having a second control-voltage terminal;

means, being connected between said first lower-voltage terminal and an electrical ground, for decoupling said first and second solid-state amplifying devices; and

said means for decoupling comprises an rf effective series resistance of less than 0.4 divided by an rf output, in Watts, of said rf power amplifier.

9. An rf power amplifier as claimed in Claim 8 in which said means for decoupling comprises a plurality of parallel-connected capacitors.

10. An rf power amplifier as claimed in Claim 8 in which said means for decoupling comprises a plurality of porcelain capacitors that are connected in parallel.

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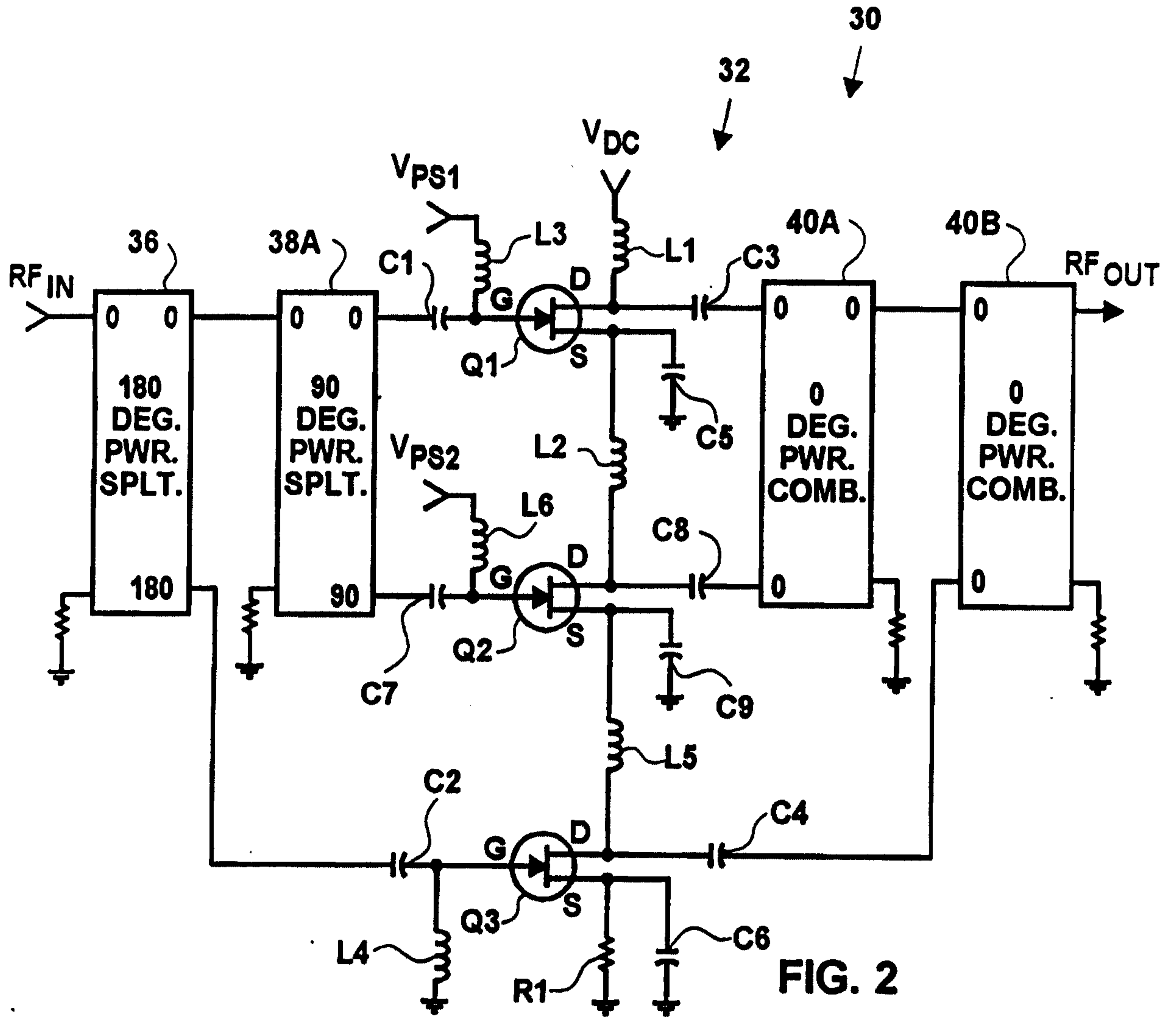


FIG. 2

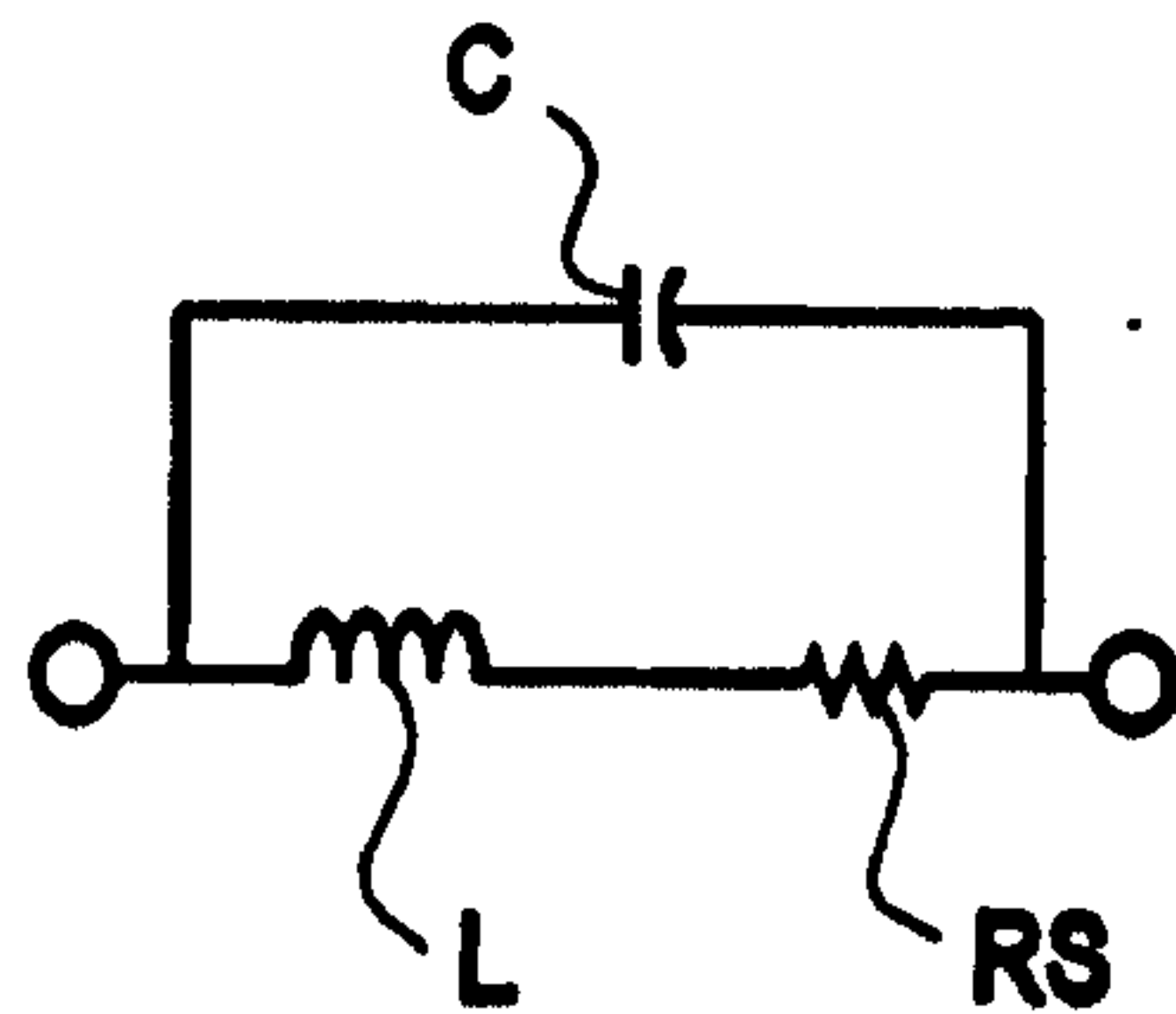


FIG. 6

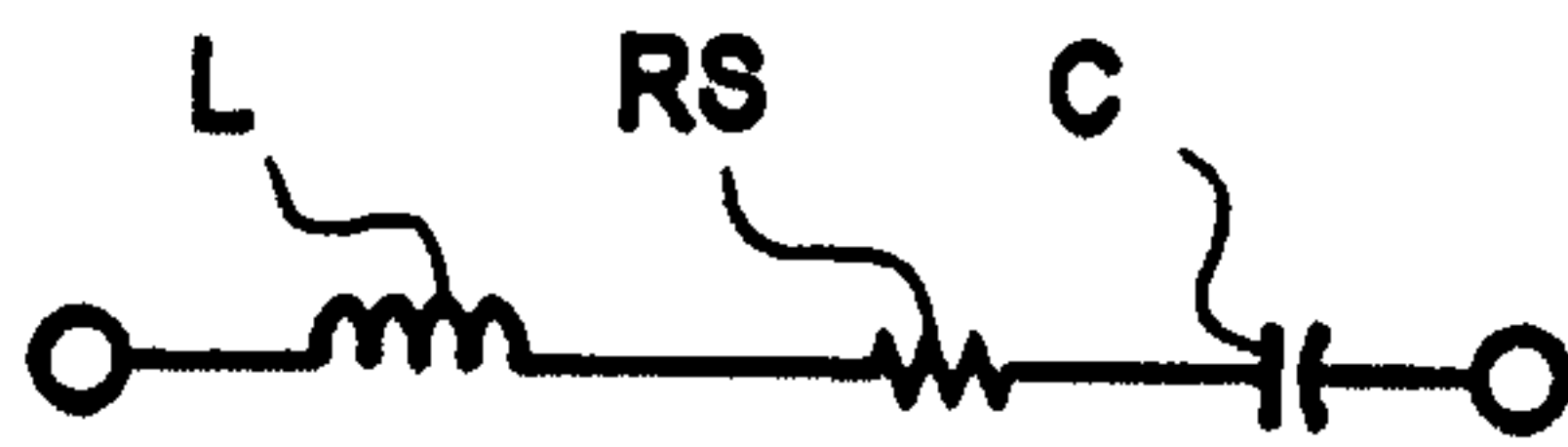


FIG. 7

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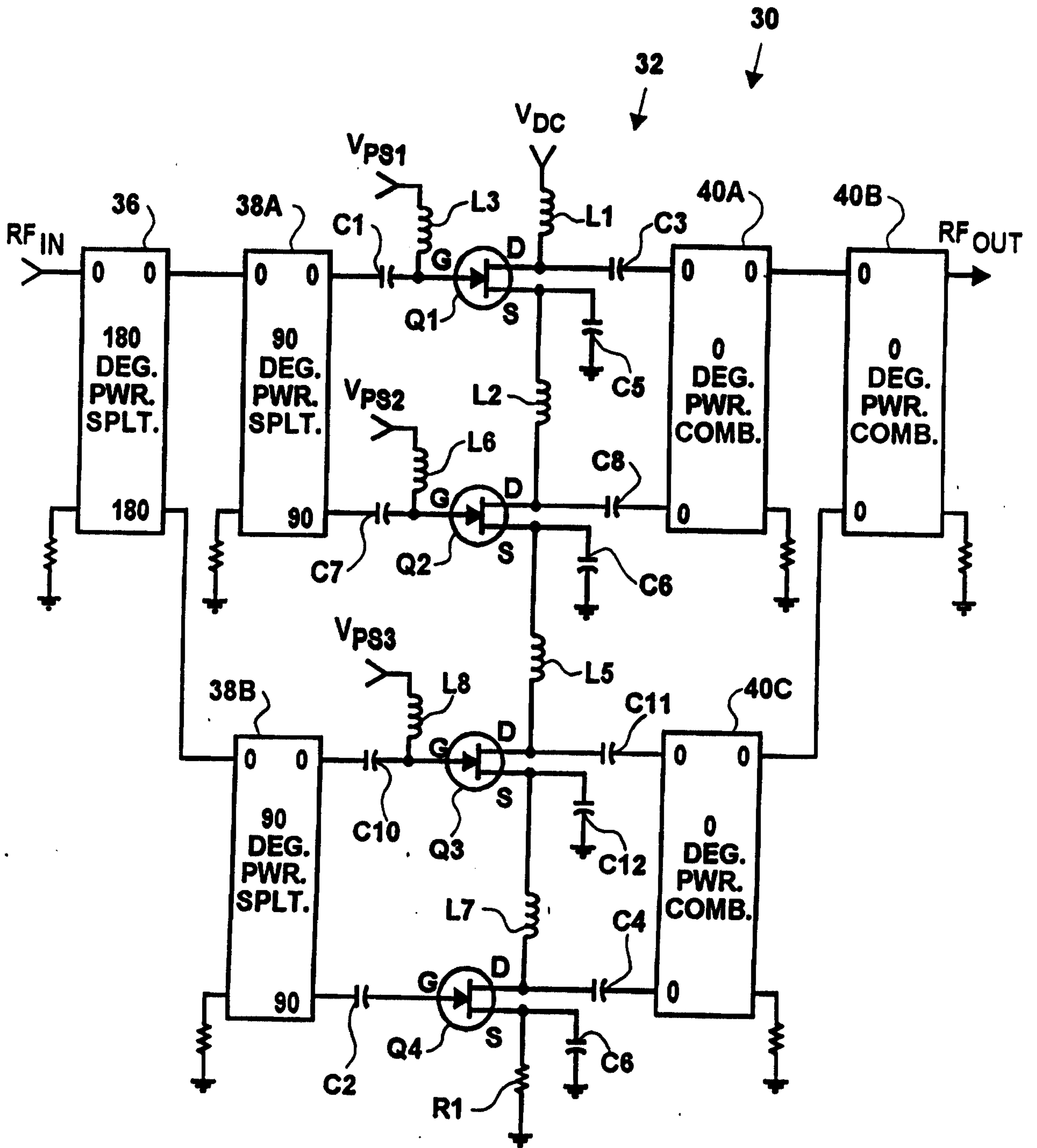


FIG. 4

