Novel RF MEMS–Enabled Circuits

4.1 Introduction

For the first time in recent history, a technology is emerging that promises to enable both new paradigms in RF circuits and systems topologies and architectures as well as unprecedented levels of performance and economy. RF MEMS is widely believed to be such a technology [1–3]. There are at least two fundamental approaches for exploiting RF MEMS, namely, bottoms-up and top-down. In the bottoms-up approach, the program to be followed would involve the direct replacement of conventional circuit elements by their superior RF MEMS counterparts, with minimum or no change in circuit topology and system architecture. In the top-down approach, the designer would begin with the proverbial clean sheet of paper, unprejudiced by the limitations and constraints imposed by conventional RF technologies, and devise circuits and systems that exploit, in the fullest possible fashion, the flexibility and power of RF MEMS. In this chapter, we present a sample of novel RF MEMS–enabled functions, which are the first fruits of both approaches. The key words describing these functions are reconfigurable and programmable (i.e., these elements, endowed by RF MEMS with the ability to be actuated, embody not just a single value, but a range of commandable values). These properties, in turn, will be exhibited by the circuits and systems of which they are a part. Since most of these visionary ideas and techniques have, understandably, anticipated RF MEMS technology maturation, the presentation will be largely of a descriptive and qualitative nature.
Nevertheless, it is expected that the ideas we have chosen to present will form the pillars upon which the paradigm of ubiquitous connectivity will be built. In order to emphasize the various ideas, the chapter is organized into three main sections dealing with circuit elements (devices), circuits, and systems.

### 4.2 Reconfigurable Circuit Elements

At the fundamental level, passive RF/microwave circuits and systems consist of switches, capacitors, inductors, transmission lines, and resonators. These devices/circuit elements were the first on which the impact and implications of RF MEMS were studied. The MEM switch, in particular, due to its noninvasive properties (i.e., virtually ideal insertion loss and isolation [1]), may be considered the most fundamental building block, or single device, embodying reconfigurability. Thus, all other reconfigurable functions will simply embody an appropriately organized set of interconnected switches. In what follows, beginning with a function employing a single switch, we expose some of the ideas elicited by the study of MEMS-based reconfigurability.

#### 4.2.1 The Resonant MEMS Switch

One of the key characteristics of RF MEMS switches is their ability to maintain good isolation over wide bandwidths (e.g., 35 dB at 40 GHz for CPW membrane MEMS switches [4]). These capacitive switches, however, exhibit poor isolation at low frequencies because in that regime the impedance of the membrane-to-center-conductor capacitance is not low enough. To overcome this limitation in this type of switch, Peroulis et al. [5] proposed the resonant MEMS switch, as shown in Figure 4.1.

The resonant MEMS switch [5] emerged from the observation that the parasitic inductance, in series with the membrane/plate capacitance, which derives from the supporting structure itself, narrows the bandwidth while enhancing the isolation of the switch. Peroulis et al. [5] exploited this realization by purposefully introducing inductive connecting beams to link the center capacitor plate to the rest of the structure (Figure 4.1). By designing the connecting inductance so as to resonate the plate capacitance at a certain frequency, the above-mentioned impedance is minimized and, consequently, the isolation is increased at that frequency. The frequency of highest isolation becomes
\[ f_R = \frac{1}{2\pi \sqrt{C_D L}} \]  

(4.1)

where \( C_D \) is the plate capacitance in the down state, given by

\[ C_D = \varepsilon_0 \varepsilon_r \frac{A}{t_d + t_{\text{roughness}}} \]  

(4.2)

where \( A \) is the area of overlap of the bridge with the center conductor of the CPW line, \( t_d \) is the thickness of the dielectric protecting the bottom electrode, \( \varepsilon_r \) is its relative dielectric constant. \( L \) is the parasitic inductance, which
can be determined either from measurements or from a full-wave electromagnetic simulation of the structure and has typical values of 0.2 to 6 pH [6]. In demonstrating the concept, Peroulis et al. [5] were able to show an isolation in the down state as high as approximately 24 dB at 13.2 GHz, with inductive connecting beams of 50 pF, in a switch 3-μm air-gap structure, over a 40/60/40-μm CPW line, with a 1,000Å-thick Si₃N₄ insulating layer.

### 4.2.2 Capacitors

Variable capacitors, or varactors, were discussed at length in Chapter 3. In this chapter, we revisit the subject of varactors, but in the context of implementations that emphasize their application in circuits other than voltage-controlled oscillators (e.g., filters).

#### 4.2.2.1 The Binary Capacitor

The binary capacitor function, a capacitance that is made to change between two values, is embodied by the shunt capacitive MEM switch. Indeed, as is well known [7], the operation of the shunt capacitive MEM switch is predicated upon the fact that, when it is in the up state, the capacitance from the bridge to the center conductor of the CPW line is very small; whereas when it is in the down state, that capacitance is very large. Thus, when examined, not from the insertion loss/isolation perspective, but from the perspective of the equivalent RF behavior of the structure, it may be readily characterized as a binary capacitor and exploited for tuning/reconfigurability purposes. Recently, Peroulis et al. [8] employed the serpentine-spring-supported low-voltage capacitive MEM switch [9] to demonstrate just such a binary capacitor (Figure 4.2). The structure’s intrinsic capacitance is given by

\[
C_p = \begin{cases} 
\frac{\varepsilon_0 A}{d + \frac{t_d}{\varepsilon_r}} + C_{\text{fringing}}; & \text{up state} \\
\frac{\varepsilon_0 \varepsilon_r A}{t_d + t_{\text{roughness}}}; & \text{down state}
\end{cases}
\]  

(4.3)

where \(A\) is the area of overlap of the bridge with the center conductor of the CPW line, \(d\) is the bridge-to-substrate distance, \(t_d\) is the thickness of the dielectric protecting the bottom electrode, \(\varepsilon_r\) is its relative dielectric constant,
Figure 4.2  (a) Capacitive MEMS switch employed as a binary capacitor over a CPW line, and (b) binary capacitor equivalent circuit model. (Source: [8] ©2001 IEEE.)

and $C_{fringing}$ is the fringing capacitance. Depending on the particulars of the implementation (e.g., Figure 4.2) and the intended application, however, it is worth pointing out that there may be other parasitic elements that would blur the simple capacitive behavior implied by (4.3), and thus they must be included in its model. For example, Peroulis et al. [8] found that for applications of the binary capacitor in filters, the extra elements in Figure 4.2(b) must be included. These elements include (1) the resistance and inductance of the bridge, denoted $R_p$ and $L_p$, respectively, and obtainable from full-wave simulations; (2) the loss in the dielectric insulator coating the bottom electrode, $G_p$, and given by [8]

$$G_p = \begin{cases} 0 & \text{up state} \\ \omega C_p \tan \delta & \text{down state} \end{cases}$$

(4.4)

where $\tan \delta$ is the loss tangent of the dielectric; and (3) the short access lines of length $l_L$ and $l_R$, as represented by their effective series resistance and inductance, $R_{SL}$ and $L_{SL}$, respectively. Peroulis et al. [8] pointed out that the parasitic inductance is particularly important in the model because its values are comparable to those found in actual filters. Under the short line approximation, assumed by Peroulis et al. [8], the parasitic elements are given by

$$R_{SL} = 2\alpha Z_0 l_L \quad \text{and} \quad R_{SR} = 2\alpha Z_0 l_R$$

(4.5)
where $\alpha$ represents the line attenuation constant, usually obtained experimentally, and

$$L_{SL} \equiv \frac{2Z_0}{\omega} \tan \left( \frac{\beta l_L}{2} \right) \equiv \frac{Z_0 \beta l_L}{\omega} \quad \text{and}$$

$$L_{SR} \equiv \frac{2Z_0}{\omega} \tan \left( \frac{\beta l_R}{2} \right) \equiv \frac{Z_0 \beta l_R}{\omega} \quad (4.6)$$

4.2.2.2 The Binary-Weighted Capacitor Array

In the most general case, there are at least two types of tuning needs to be considered in RF/microwave circuits and systems. First, one might be interested in changing the frequency-determining parameters of a circuit/system, designed originally to operate in a given frequency band, so that it can operate at a different frequency band. An example of this would be a filter operating in a multiband wireless handset, whose response could be switched back and forth so that it could operate in two or more frequency bands. Second, one might be interested in controlling the frequency-determining parameters of a circuit/system so as to optimize its performance, which may well include switching among bands, but more interestingly would include enabling the adaptive control of the circuit/system so that it can meet a certain error function, predicated for instance, upon the real-time insertion loss, return loss, power efficiency, or harmonic level. The binary-weighted capacitor and inductor arrays [10] shown in Figures 4.3 and 4.4 were devised with these scenarios in mind.

![Diagram](image)

**Figure 4.3** Binary-weighted capacitor array. *(After: [10].)*
In the binary-weighted capacitor array, the noninvasive properties of MEM switches are exploited to reconfigure networks of elementary unit capacitor cells (Figure 4.3) so that the overall network realizes a certain capacitance. Thus, in essence, the approach discretizes the top plate area of the overall capacitor so that any value of capacitance, in steps of the unit capacitor cell value, may be set by opening or closing appropriate switches. Thus (for example, in Figure 4.3), assuming each unit capacitor cell has a value \( C \), then to obtain an overall network capacitance of \( 2C \), the switches \( S_3 \) and \( S_4 \) would be closed, while the switches \( S_1, S_2, S_5, S_6, S_7, S_8, S_9, \) and \( S_{10} \) would be open. To obtain a network capacitance of \( 4C \), the switches \( S_6, S_7, S_8, S_9, \) and \( S_{10} \) would be closed, while the switches \( S_1, S_2, S_3, S_4, \) and \( S_5 \) would be open.

### 4.2.3 Inductors

#### 4.2.3.1 The Binary-Weighted Inductor Array

It is well known that inductors and capacitors for operation at microwave frequencies may be realized from short sections of transmission lines (e.g., of length less than one-quarter the operating wavelength) \([11]\). Thus, a tunable inductor may be obtained by interconnecting, preferably via MEMS switches, network unit inductor cells so that the overall network exhibits a certain inductance. Such is the idea behind the binary-weighted inductor array \([10]\) shown in Figure 4.4. For example, in order to obtain an inductance value of \( L_{21} \), switches \( S_1, S_2, \) and \( S_3 \) would be closed, and switches \( S_4, S_5, S_6, S_7, S_8, S_9, S_{10}, S_{11}, \) and \( S_{12} \) would be open. On the other hand, if switches with an inductance of value \( 2L_1 + 2L_2 \) were desired, then switches \( S_1, S_3, S_4, S_5, S_7, S_{10}, \) and \( S_9 \) would be closed, while switches \( S_2, S_6, S_{11}, \) and \( S_{12} \) would be open.
4.2.3.2 Series and Shunt Tunable Inductor Arrays

In addition to the systematic (binary-weighted) inductor array considered above, other arrangements for tuning series and shunt inductors, in the context of MEMS switch switching, have been advanced, as shown in Figures 4.5 to 4.7 [12]. In Figure 4.5, MEMS switches $S_1$, $S_2$, ..., $S_x$ are connected in parallel with inductors $L_1$, $L_2$, ..., $L_x$, which, in turn, are connected in series between nodes IN1 and OUT1. With all switches open, the total inductance between nodes IN1 and OUT1 equals the sum of all the series-connected inductors. When any of the switches is closed, however, the low impedance of the switch bypasses that of the inductor, in essence short-circuiting it, and the total inductance decreases by that of the shorted inductor. For example, if all inductors have a common value $L$, then the total inductance between nodes IN1 and OUT1 may be set to be any multiple of $L$ from a minimum of $L$ to a maximum of $X$ times $L$ (i.e., $XL$), where the minimum value is obtained by closing all switches but one, and the maximum value is obtained by opening all switches. If all switches are closed, a nearly zero inductance is

![Diagram of series-connected shunt-switched inductor array and parallel-connected series-switched inductor array. (After: [12].)
obtained. In Figure 4.6, MEMS switches $S_1$, $S_2$, ..., $S_x$ are connected in series with inductors $L_1$, $L_2$, ..., $L_x$, which, in turn, are connected in parallel between nodes IN2 and OUT2. When all switches are closed, the reciprocal of the total inductance, between nodes IN2 and OUT2, equals the sum of the reciprocals of all the parallel-connected inductors. When any of the switches is open, however, the high impedance of the switch essentially disconnects the inductor in question. For example, in a configuration of four inductors with a common value $L$, the total inductance will vary from a maximum of $L$, when all switches are open but one, to a minimum value of $L/4$, when all switches are closed. Clearly, by combining the series- and parallel-connected inductor arrays, an even more ample and fine-grained set of inductance values becomes available (Figure 4.7).

### 4.2.4 Tunable CPW Resonator

It is well known that one of the factors limiting the self-resonance frequency of spiral inductors is the inevitable need to use an air-bridge for connecting the inner terminal of the spiral to the output terminal, and the concomitant parasitic capacitance between the air-bridge and the underlying spiral traces. The tunable CPW resonator, demonstrated by Ketterl, Weller, and Fries [13], exploits this topological accident in the spiral inductor in order to transform it into a tunable resonator (Figure 4.8).
In the tunable resonator, the air-bridge is replaced by an electrostatically actuated cantilever beam anchored at the center of the spiral. The spiral inductor, together with the cantilever beam-to-spiral capacitance, embodies an \(LC\) resonator. When a voltage beyond pull-in is applied between the cantilever beam and the underlying SiO-coated spiral inductor, the beam “zipping” deflection action changes the beam-to-spiral capacitance, thereby changing the resonance frequency of the structure. A prototype achieved resonance tuning between 4 and 7 GHz, under biases from 0 to 40V, with corresponding \(Q_s\) between 17 and 20.

### 4.2.5 MEMS Microswitch Arrays

So far in this chapter we have addressed techniques for the reconfigurability of more or less discrete circuit elements. Another vein in the area of
Figure 4.8 Tunable CPW resonator: (a) layout; (b) optical image of fabricated device; and (c) side view of suspended beam showing the cantilever. Top: at 0V bias; middle: after pull-in; bottom: in continual tuning state with increase in bias as the beam zips. (Source: [13] ©2001 IEEE. Courtesy of Prof. T. Weller and Mr. T. Ketterl, University of South Florida.)
reconfigurability—namely, the MEMS microswitch array [14]—addresses the reconfigurability of distributed microwave components (i.e., of the very metal traces, or patterns, that would otherwise define the interconnection transmission lines and tuning stubs of microstrip-based microwave circuits). The fundamental enabler of this paradigm, the microswitch, is shown in Figure 4.9. The microswitch is a cantilever beam–type structure that can be arrayed in two dimensions with an interelement pitch of 100 μm. For implementation on a fused silica substrate ($\varepsilon_r = 3.8$), this size corresponds to $1/20$th the wavelength at 100 GHz or $1/200$th the wavelength at 10 GHz, so that no issues of line-length quantization are elicited [14]. By addressing the two-dimensional array, where each microswitch may be thought of as a pixel, any given metal pattern image can be defined on the substrate, particularly as it is appropriate to a matching or tuning network. For example, Figure 4.10 shows the microswitch array-based tuning for a reconfigurable power amplifier, in which both the input and output matching networks are reconfigured to retune the optimum frequency of the amplifier.

### 4.3 Reconfigurable Circuits

Impedance matching is one of the fundamental steps in the design and production of an RF/microwave circuit [15, 16]. In low-noise amplifiers (LNAs) and power amplifiers, properly tuned input/output matching networks are

![Figure 4.9](image.png)  
**Figure 4.9** Conceptual diagram of the 2-D microswitch array. Inset: SEM of a single microswitch. The corrugations are added for mechanical strength in the cantilever arm. (*Source:* [14] ©2000 IEEE.)
crucial to meeting required noise figure and power efficiency requirements. In the production of low-volume MICs, or hybrid (discrete) circuits, it is the rule to manually tune the circuits until the desired performance levels are met; but this activity becomes too time consuming and expensive at millimeter-wave frequencies, not to mention impractical for high-volume applications and even impossible for MMICs. Thus, there is a strong incentive to exploit the power of RF MEMS to implement classic impedance matching schemes in an automated, reconfigurable fashion.

4.3.1 Double-Stub Tuner

A case in point that pertains to reconfigurable impedance matching is the demonstration by Lange et al. [17] of a reconfigurable double-stub tuner using MEMS switches. The double-stub tuner [18] is a popular impedance matching technique in microwave circuit design. Given a load, $Y_L = G_L + jB_L$, the technique exploits the impedance-transforming properties of transmission lines to transform its real and imaginary parts, $G_L \rightarrow G_m = Y_0$ and $B_L \rightarrow$
\( B_{in} = 0 \), into a real input impedance, \( Y_{in} = G_{in} = Y_0 \), which represents a perfect match if \( Y_0 \) is the characteristic admittance of the system. This is accomplished via the circuit topology of Figure 4.11, whose input admittance is given by [19]

\[
Y_{in} = Y_0 \frac{G_L + jB_L + jB_1 + jY_0 \tan \beta d}{Y_0 + (G_L + jB_L + jB) \times j \tan \beta d} + jB_2 \quad (4.7)
\]

By solving for the real and imaginary parts of (4.7) and imposing conditions on \( B_1, B_2, \) and \( d \), so that \( G_{in} = Y_0 \), and \( B_{in} = 0 \), the matching problem is solved. In particular [17], the stub separation distance \( d \) is chosen such that

\[
0 \leq G_L \leq \frac{Y_0}{\sin^2 \beta d} \quad (4.8)
\]

and \( B_1 \) and \( B_2 \) are given by

\[
B_1 = -B_L \pm \frac{Y_0 + \sqrt{(1 + \tan^2 \beta d)G_L Y_0 - G_L^2} \tan^2 \beta d}{\tan \beta d} \quad (4.9)
\]

and

\[
B_2 = \frac{\pm Y_0 + \sqrt{(1 + \tan^2 \beta d)G_L Y_0 - G_L^2} \tan^2 \beta d}{G_L \tan \beta d} + G_L Y_0 \quad (4.10)
\]

\[ Y_v \] and \[ Y_t \] are the input and output admittances, respectively, and \( G \) represents the conductance of the stubs. The diagram shows the double-stub tuner topology.

**Figure 4.11** Double-stub tuner topology.
where $\beta$ is the phase constant in the transmission line. Equation (4.7) imposes a limitation on the possible values of load admittance $G_L$ that may be matched in terms of $d$, wavelength $\lambda$, and $Y_0$. For example, for $d = 0.1\lambda$ and $Y_0 = 0.02S$, $GL$ must be less than 0.057$S$ in order to be matched to 0.02$S$ (50 ohms).

The reconfigurability in the approach of Lange et al. [17] (Figure 4.12) derives from their use of RF MEMS switches to switch in and out multiple parallel-connected stubs, thus opening up the potential for realizing multiple values of $B_1$ and $B_2$. This, in turn, greatly extends the range of possible load impedance values that can be matched. With reference to Figure 4.12, then, when all switches are off, the open-circuited stubs (represented by $C_{\text{fixed}}$) would each adopt an equivalent capacitance given by

$$C_{\text{eq}} = \frac{C_{\text{fixed}} \times C_{\text{off}}}{C_{\text{fixed}} + C_{\text{off}}} = C_{\text{on}}$$  \hspace{1cm} (4.11)

if $C_{\text{on}} \ll C_{\text{fixed}}$, thus effectively disconnecting the stub from the common node $N$. If the switch is on, however, then, if $C_{\text{on}} \gg C_{\text{fixed}}$ the equivalent stub capacitance is given by

$$C_{\text{eq}} = \frac{C_{\text{fixed}} \times C_{\text{on}}}{C_{\text{fixed}} + C_{\text{on}}} = C_{\text{fixed}}$$  \hspace{1cm} (4.12)

The stub capacitance required to fulfill (4.11) and (4.12) is a function of the frequency of operation and the values of susceptance, $B_1$ and $B_2$, and is given by [17]

![Figure 4.12 Reconfigurable stub. (After: [17].)]
\[ C_{\text{fixed}} = \frac{B}{2\pi f} \] (4.13)

For the 4-bit reconfigurable stub prototype demonstrated by Lange et al. [17], which matched loads with real and imaginary parts between 20 and 80Ω and −150 to +150Ω, respectively, the stub capacitor values ranged between 45 fF and 1,155 fF, and the switches possessed capacitances \( C_{\text{switch}}^{\text{off}} \approx 35\ fF \) and \( C_{\text{switch}}^{\text{on}} \approx 3\ pF \). With each 4-bit stub capable of adopting 16 susceptance values, a total of 256 configurations, or the equivalent of 256 double-stub tuner realizations, are possible—indeed, a powerful testimony to the power of RF MEMS and reconfigurability.

4.3.2 \textit{Nth}-Stub Tuner

The double-stub tuning technique described above relies on three parameters: the susceptances, \( B_1 \) and \( B_2 \), of two shunt stubs, and the distance separating them, to match a load to the working characteristic impedance of the system, usually 50Ω. Its disadvantages are that the range of loads that can be matches is limited and that it is narrowband. The triple-stub tuner system [18], in turn, can match all values of load admittances and can be optimized to increase the bandwidth. The stub-tuning concept may be generalized to that shown in Figure 4.13 [10], where a transmission line of a certain length, demarcated by its input and output ports, could have connected to it via MEM switches one or more shunt stubs of predetermined lengths at selected locations along the transmission line. By selectively closing MEM switches to some of the shunt stubs while opening others, a desired frequency response for the transmission line can be obtained to effect impedance matching within the desired frequency range. The spacing for adjacent shunt stubs along the transmission line is about one-quarter wavelength, or an integral multiple of

\[ \text{Figure 4.13} \quad \text{Nth-stub/programmable transmission line tuner. (After: [10].)} \]
one-quarter wavelength. The length of each shunt stub is preferably about half a wavelength, or an integral multiple of half a wavelength. However, many combinations of shunt stub lengths and spacing are also possible.

An example of an application where the $N$th-stub tuner offers an enabling function is the harmonically tunable power amplifier (Figure 4.14) [10]. Due to the high power levels produced, these power amplifiers exhibit a nonlinear transfer response characteristic that produces undesired harmonic output signals. The harmonic signals are generated at frequencies that are integral multiples of the fundamental frequency (i.e., the frequency of the input signal). With programmable stub-tuning transmission lines, to affect input and output impedance matching, a programmable stub-tuning transmission line that branches off from the output of the amplifier is added to reduce the harmonics generated by the amplifier. The harmonic tuning transmission line is connected to a plurality of shunt stubs via respective MEM switches, which are selectively turned on or off to reduce the amplitudes of harmonic signals. In general, an amplifier may produce several harmonic frequency signals that must be suppressed, and the shunt stubs with respective terminations, which may be opens or short-circuits, are arranged to prevent these signals from being transmitted to the output Out. A properly tuned harmonic tuning line segment directs the harmonic signals to a

![Harmonically tunable power amplifier diagram](image)

**Figure 4.14** Harmonically tunable power amplifier. *(After: [10].)*
load, which absorbs the energy of the harmonic signals, so that they are not reflected back to the output line’s input X. The shunt stubs with different spacing or different lengths may be used to optimize the absorption of the harmonics of interest.

### 4.3.3 Filters

Filters are ubiquitous building blocks in wireless systems. Indeed, a great deal of time is spent designing and redesigning filters (e.g., in satellite communications, where each satellite program must operate at a distinct set of frequencies). Reconfigurable filters, therefore, would result in more economical wireless systems, not only because fewer filters would have to be realized, but because those realized could be tuned, in principle, by computer means, as opposed to manually. Figure 4.15 shows a possible application of the binary-weighted capacitor and inductor arrays described previously to programmable filters. In the topology shown in Figure 4.15(a), an input transmission line is connected to a fix capacitor C1, which, in turn, is connected to a binary-weighted capacitor array C1a, where C1a includes a plurality of

![Filter Circuit Diagram](image-url)

**Figure 4.15** Programmable microwave filter. *(After: [10].)*
MEM switches to adjust the network’s capacitance. The capacitor \( C_1 + C_{1a} \), in turn, is connected to a binary-weighted inductor circuit \( L_1 \). A filter is formed by repeating the structure of alternately connected capacitors \( C_2 + C_{2b}, C_3 + C_{3c}, 22a \), and the inductor network \( L_2 \). The filter’s output signal is transmitted from the capacitor \( C_3 \) to the output transmission line. Figure 4.15(b) shows the equivalent circuit of Figure 4.15(a), in which the capacitances \( C_{1a}, C_{1b}, \) and \( C_{1c} \) are the sums of the capacitance values of the capacitors \( C_1, C_2, \) and \( C_3 \) and the tunable capacitor networks \( C_{1a}, C_{2b}, \) and \( C_{3c} \), respectively. The inductances \( L_a \) and \( L_b \) are the inductance values of the tunable inductive line networks \( L_1 \) and \( L_2 \), respectively. Each of the capacitance values \( C_{1a}, C_{1b}, \) and \( C_{1c} \) and the inductance values \( L_a \) and \( L_b \), which together determine the frequency response of the filter, can be changed by selectively switching at least some of the MEM switches within the capacitor and inductor networks.

### 4.3.4 Resonator Tuning System

Many modern wireless systems use resonator tuning schemes for changing communication frequencies. Most methods for changing communication frequencies are predicated upon coupling a voltage-controlled capacitor (varactor) to a resonator in order to change its resonance frequency. The tunable CPW resonator [13] discussed in Section 4.2.4 is an example of such a scheme, although in that case the varactor, implemented with a cantilever beam, is both the capacitor of the \( LC \) resonator and the means to effect frequency tuning. The fundamental disadvantage of varactor-coupled tuning approaches is that the intrinsic parasitic resistance of the varactor introduces losses in the resonator, thus lowering its unloaded \( Q \). The consequence of a reduction in the unloaded \( Q \) may be appreciated by examining the carrier-to-noise (C/N) ratio in a voltage-controlled oscillator (VCO), where C/N is given by [20]

\[
\frac{C}{N} = \frac{(2 \times Q_L \times \Delta f)^2 \times P_0}{(\text{Loss} \times f_0)^2 \times (2 \times kT \times B \times NF)} \quad (4.14)
\]

where \( Q_L \) is the loaded \( Q \) of the resonator, Loss is the loss factor in the resonator, \( f_0 \) is the frequency of oscillation, \( \Delta f \) is the offset frequency from \( f_0 \), \( P_0 \) is the output power of the oscillator, \( k \) is Boltzmann’s constant, \( T \) is absolute temperature, \( B \) is the measurement bandwidth, and \( NF \) is the noise figure of the amplifier. Examination of (4.14) reveals that in order to obtain high C/N
ratio, the loaded $Q$ must be high. The loaded $Q$, in turn, is highest when the resonator experiences minimum external loading.

A novel technique to effect resonator tuning, which is enabled by an electrostatically actuated MEM air bridge, is indicated in Figure 4.16. In this scheme, changing the resonator’s resonance frequency is accomplished by varying the capacitor or varactor coupling, rather than by varying the capacitor [21]. In essence, an interferometer, such as a Mach-Zender interferometer [19], is coupled to the resonator. Then, by way of an electrostatically actuated air-bridge disposed over one of its arms, its transmission, and consequently its coupling to the resonator, changes the resonance frequency of the resonator as described below. Using impedance-transforming properties of a transformer, the input to the primary port of a $1:N_r$ transformer whose secondary is loaded with a capacitor $C_{\text{tuning}}$ or an inductor $L_i$, results in a capacitance $N_r^2 C_{\text{tuning}}$ or an inductance $L_i/N_r^2$, respectively. To vary the effective coupling $N_r$, a Mach-Zender interferometer is coupled to dielectric resonator. In Figure 4.16, the Mach-Zender interferometer is implemented as a capacitor, specifically a ring capacitor. Thus, the Mach-Zender interferometer acts as a tunable capacitor and includes a bottom electrode, an air bridge, and a ring branch. Applying an actuation voltage $V_i$ causes the air bridge to

![Figure 4.16](image)

**Figure 4.16** Coupling-based resonator tuning. (*After:* [21].)
deflect towards the ring branch, thus loading the ring branch with variable
capacitance, which, in turn, changes the coupling to the ring branch and, as a
consequence, the effective capacitance coupled to resonator. The dielectric
resonator is coupled to a transmission line having a termination \( R_T \) at one
end and an active element at the opposite end. The active element is coupled
to both a feedback element and a matching network that is coupled to a ter-
minating load \( R_L \).

Referring to Figure 4.17, assuming balance amplitudes (i.e., \( |t_1| = |t_2| = 1 \)), the transmission \( T \), which relates the output-to-input wave amplitude
ratio, is given by

\[
T = |t|^2 = 2 \times \left( 1 + \cos(L_\pi (k_2 - k_1)) \right)
\]  

(4.15)

where \( k \) is the propagation constant defined by

\[
k = \omega \sqrt{L/C}
\]

(4.16)

with \( \omega \) being the frequency and \( L \) and \( C \), being the inductance and capaci-
tance per unit length, respectively. \( L \) is one-half the mean circumference of
the ring. For a given \( L \), \( T \) is a function of \( k_1 \) and \( k_2 \), and \( T \) is a measure of the
coupling between input and output when there is an output transmission
line. When there is not an output transmission line, the waves in each branch
of the ring simply counter-propagate and \( T \) still represents the coupling to

\[\begin{array}{c}
\text{In} \\
\text{Out}
\end{array}\]

\[
t_1 \sim e^{\frac{\pi L}{k_1}} \\
t_2 \sim e^{\frac{\pi L}{k_2}}
\]

Figure 4.17  Mach-Zender interferometer/ring capacitor. (After: [21].)
the ring. In this case, however, it is more appropriate to consider the reflection from ring \( R = 1 - T \). This coupling can be varied by changing \( k_2 - k_1 \), in particular, by changing \( C \), on one of the ring branches. Thus, the concept allows the tuning of the resonator yet without deteriorating its \( Q \) by the tuning mechanism.

### 4.3.5 Massively Parallel Switchable RF Front Ends

Many communications systems demand the ability to receive narrowband signals that can occur anywhere, in any one of a number of channels, within a wide frequency band. Since the equipment receiving these signals must operate in coexistence with high-power transmitters [22] and because it is imperative to avoid interference, it is required that excellent filters with narrow instantaneous bandwidth, high out-of-band rejection, wide tunability, and low insertion loss be utilized. As these requirements are only met by unrealizable filters, the usual approach to the problem involves the parallelization of the receiver into independent channels, each one containing a filter of realizable characteristics (Figure 4.18) [22]. The key to this massively parallel receiver scheme, however, is to utilize, for signal routing and reconfiguration purposes, RF/microwave switches with virtually ideal performance (particularly low insertion loss) because they directly impact the noise figure,
high linearity (because in some case they must route high-power signals), and low power consumption (because of the large number of them that are required). This is precisely one of the quintessential opportunities that may be enabled by RF MEMS switches [22]. Brown [22] pointed out that traditional implementations of this architecture, based on conventional switch technology (i.e., pin diode switches), are very massive, power-consuming, and expensive. For example [22], if the pin diode switches utilized in the front-end of the ARC-210—perhaps the premier radio for military airborne communications in the VHF and UHF bands between 30 and 400 MHz—were to be replaced by MEM switches, the front-end noise figure would improve by 0.5 dB (from 4.5 to 4 dB), the transmitter-to-receiver isolation due to MEM switches in combination would improve by 20 dB (from 60 to more than 80 dB), and the total power consumption would be reduced from approximately 100 mW to less than 1 mW. Clearly, these projections could apply to the multiband/multistandard wireless transceivers that will enable the ubiquitous communications vision.

### 4.3.6 True Time-Delay Digital Phase Shifters

Phase shifters are at the heart of phased array antennas [23]. In simple terms, a phased array antenna consists of a set of phase shifters that control the amplitude and phase of the excitation to an array of antenna elements in order to set the beam phase front in a desired direction. While phase shifters that provide a lumped-element circuit phase shift as well as those providing a physical time delay phase shift may be employed in the implementation of a phased array, the true time delay approach enables frequency-independent beam steering, which permits the realization of phased arrays with wide instantaneous bandwidth [22]—a highly desirable feature. Consider, for example, the conventional and true time delay phased array antenna schemes depicted in Figure 4.19.

In both schemes, we find two antenna elements separated by a distance $d$, and driven through phase shifters in such a way that beams are set up in the direction $\theta_1$ when the input frequency is $\omega_1$, and $\theta_2$ when the input frequency is $\omega_2$. We notice, however, that whereas in Figure 4.19(a) the beam direction when the input frequency is $\omega_2$ differs markedly from that when the frequency is $\omega_1$, in Figure 4.19(b) the beam direction for the two frequencies is virtually identical. Let us examine this situation.

To maximize radiation in the direction $\theta_1$, the waves emitted from the adjacent antenna elements must interfere constructively in that direction, which requires that the path length difference between these waves (namely, $k_1$, $k_2$, and $k_3$)
\[ d \sin \theta \]

\[ \Delta \phi \]

Thus, we obtain the relation \( \theta = \sin^{-1}(\Delta \phi \times c/\omega d) \), which gives the beam direction in terms of the difference in phase of the excitation between the two elements, the frequency, and the separation. If the phase shift between elements \( \Delta \phi \) varied linearly with frequency, then the ratio \( \Delta \phi/\omega \) would be frequency-independent, and therefore, the beam direction would be independent from input signal bandwidth. This frequency-independence of the phase shift is difficult to achieve in lumped-element LC circuits; however, it is easy to obtain with the true time delay phase shifter approach, where \( \Delta \phi = k(L_2 - L_1) = \omega(L_2 - L_1)/\nu \) and \( L \) and \( \nu \) are the physical length and the velocity of propagation in the delay lines, respectively. Inserting this value into the direction angle gives \( \theta = \sin^{-1}(c(L_2 - L_1)/\nu d) \), which is frequency-independent.

Clearly, since with a phased array antenna one is interested in directing the beam, possibly containing broadband signals, into a multitude of directions, it is necessary to employ true time delay phase shifters with not just two, but as many as practical phase shift states. One phase shift topology, which is enabled by MEM switches, is shown in Figure 4.20.

In this digital phase shifter, the overall phase shift is set by properly configuring the switches so that the RF signal is directed through one of \( 2^N \) input-to-output path lengths and binary loop combinations. Clearly, the performance of the switches, particularly their insertion loss and isolation, is critical for the successful implementation of the scheme—and what better candidate than RF MEMS switches!
Figure 4.20  Schematic diagram of true time delay phase shifter in which three different binary loops are connected in series to provide \(2^3\) possible electrical delays between input and output. Extending the topology to \(N\) stages to produce \(2^N\) possible delays is obvious. (Source: [21] ©1998 IEEE.)

4.4 Reconfigurable Antennas

As is well known, the radiation properties of antennas depend on the relationship between some characteristic length in their structure and the frequency being radiated. For example, dipole antennas are nearly resonant at a length close to one-half the wavelength of the excitation signal [18]. It is logical, then, that in order to increase the flexibility and usability of antennas, one would look into ways of reconfiguring their structure and dimensions.

4.4.1 Tunable Dipole Antennas

In the tunable dipole antenna [24] (Figure 4.21), a set of symmetrically located center-fed and segmented dipoles are networked via a two-dimensional array of MEM switches. Then, by closing and opening the MEM switches in an intelligent manner, the shape and length of the antenna are reconfigured and, consequently, its radiation pattern. The concept can easily be extended to a variety of antennas, such as Yagi-Uda antennas, log periodic antennas, helical antennas, and spiral plate and spiral slot antennas [25].

An extension of the tunable dipole antenna is the reconfigurable multiband microstrip resonator antenna, as shown in Figure 4.22 [26]. In this antenna, a microstrip resonator 16, on substrate 20, designed to radiate at the highest frequency (band) of interest, is excited by a signal traveling down the microstrip line, printed on substrate, through a coupling slot, defined on ground plane. The lengths of both the resonator and the coupling slot are chosen to be approximately one-half of the wavelength [26] corresponding to the highest frequency of interest. The antenna is made reconfigurable by MEM switch, which, when in the on state, changes the resonator length to
include the additional piece of transmission line. The physically longer resonator resonates and radiates most efficiently at a lower frequency (band). To maximize the power coupling from microstrip through slot to the resonator, the microstrip is terminated by quarter-wave open-end tuning stubs, which induce short-circuits at the junction of the highest and lowest frequencies of interest with the microstrip. Then, by exploiting the high isolation of the MEM switch in the off state, and its low insertion loss in the on state, the resonator length can be switched between its high-frequency configuration, main resonator element alone, and its low-frequency configuration, which includes the extra tuning line segment. This realizes a multifrequency tunable antenna.

### 4.4.2 Tunable Microstrip Patch-Array Antennas

The tunable dipole antenna and the aperture-coupled microstrip line resonator antenna are made reconfigurable by changing the length of the radiating elements, which is directly related to the wavelength at which they resonate. Another popular radiating element, the microstrip patch, is two-dimensional in nature and finds extensive application in antenna arrays. Interest in multifrequency patch arrays led to the reconfigurable antenna array concept [27] (Figure 4.23).
In this antenna, groups of patches are electrically connected via MEM switches. The various groups operate at multiple frequencies, according to their size. By electrically connecting or disconnecting patches, the resulting overall patch geometry acquires a side effective length and shape that is concomitant with the desired frequency of operation.

### 4.5 Summary

In this chapter we have presented a mostly descriptive résumé of many novel emerging RF MEMS–enabled devices, circuits, and systems. The novel devices included the resonant MEM switch, the binary MEM capacitor, the binary-weighted capacitor and inductor arrays, the tunable CPW LC resonator, and the microswitch array. The novel circuits included the reconfigurable double-stub tuner, the Nth-stub tuner, a variety of reconfigurable filters, a resonator tuning system, as well as novel receiver and phase shifter architectures. Finally, we presented a variety of novel antenna concepts, in particular,
the tunable-dipole antenna, the slot-coupled microstrip resonator antenna, and the reconfigurable microstrip antenna. The presentation, which exposed the MEM switch, and electrostatic actuation in general, as the key elements permeating all these applications, should also serve as a resource to those individuals developing their own RF MEMS ideas and concepts. In the next chapter, we present case studies detailing the design process of RF MEMS-enabled circuits.

References


